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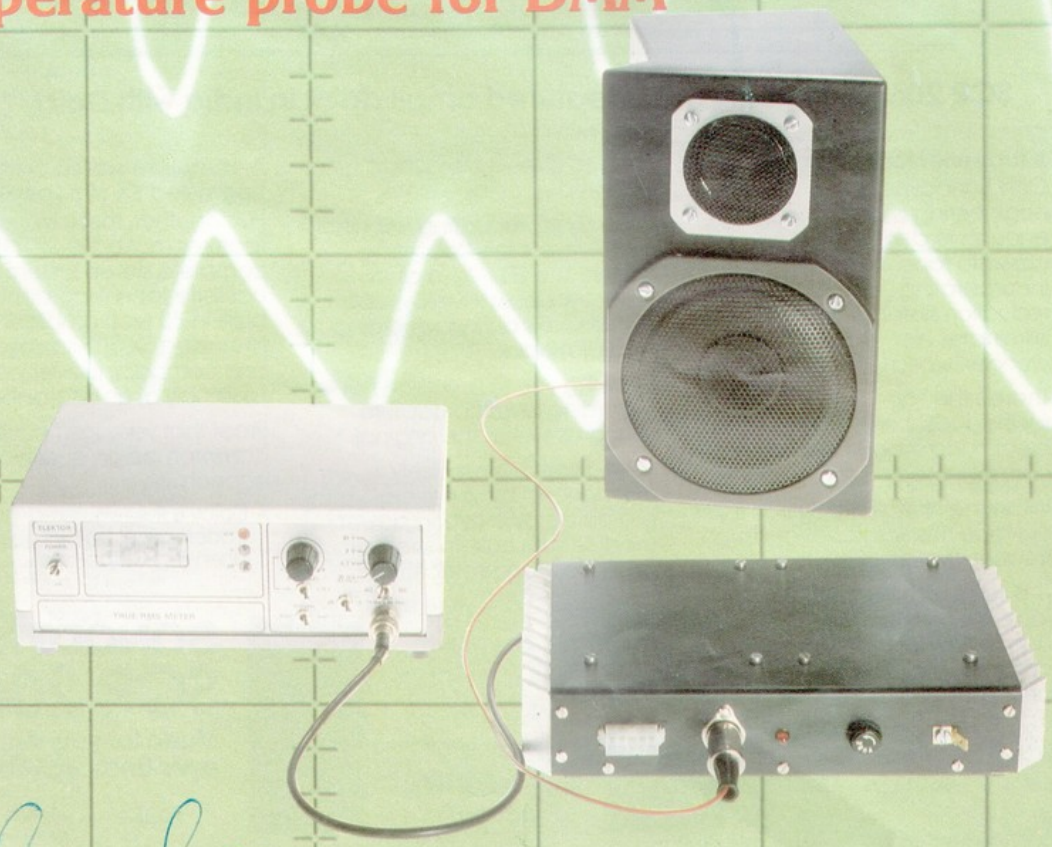

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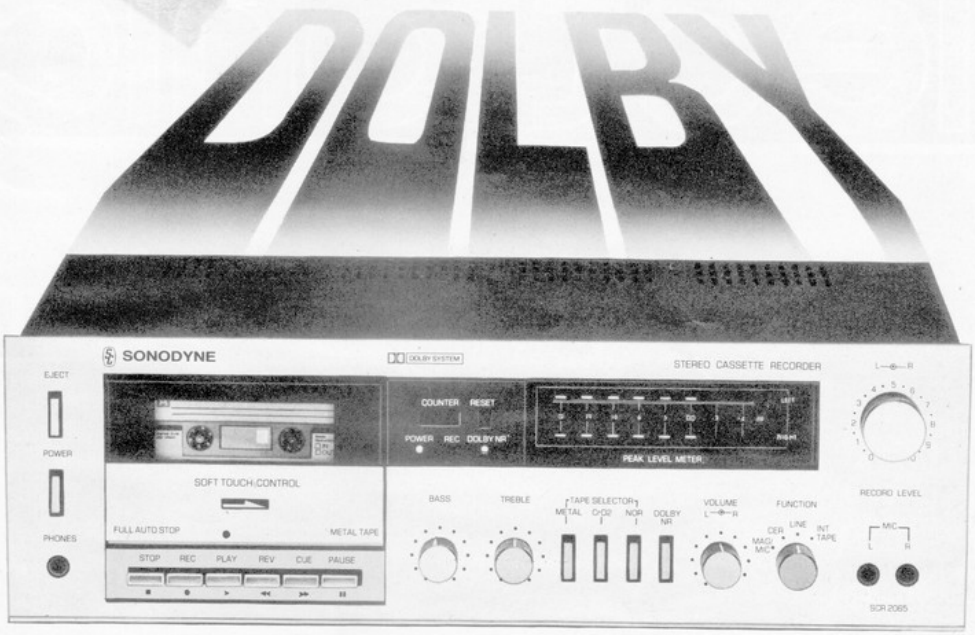

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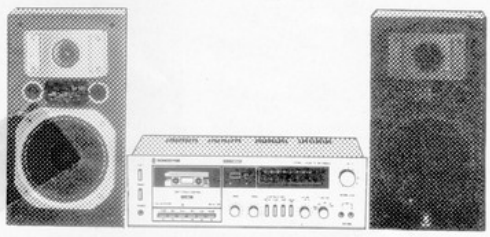

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Address:
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52, C Proctor Road
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Overseas editions:

Elektor Electronics
Standfast House
Bath Place
High Street, Barnet
Herts EN5 5XE U.K.
Editor: Len Seymour

Publitron Publicacoes Tecnicas Ltda
Av Ipiranga 1100, 9° andar
CEP 01040 Sao Paulo — Brazil
Editor: Juliano Barsali

Elektor sarl
Route Nationale; Le Seau; B.P. 53
59270 Bailleul — France
Editors: D R S Meyer;
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Elektor Verlag GmbH
Susterfeld-Straße 25
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Printed At:
Trupti Offset
Bombay - 400 013
Ph. 4923261, 4921354

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The Netherlands

elektor electronics

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January 1987

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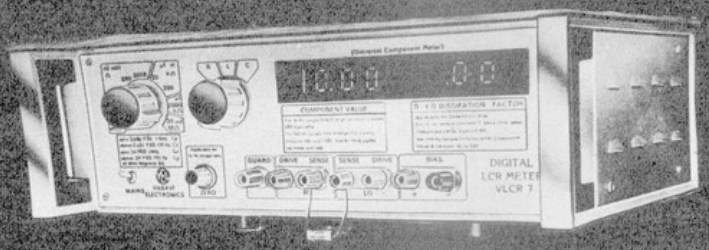

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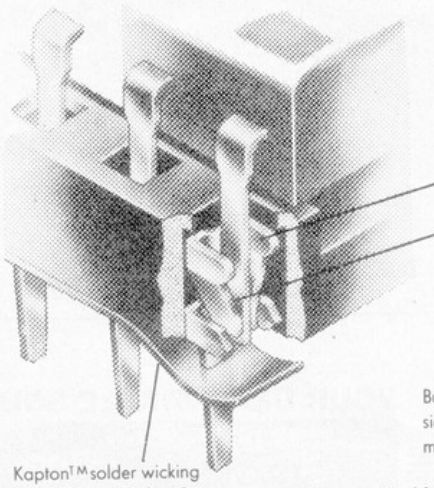

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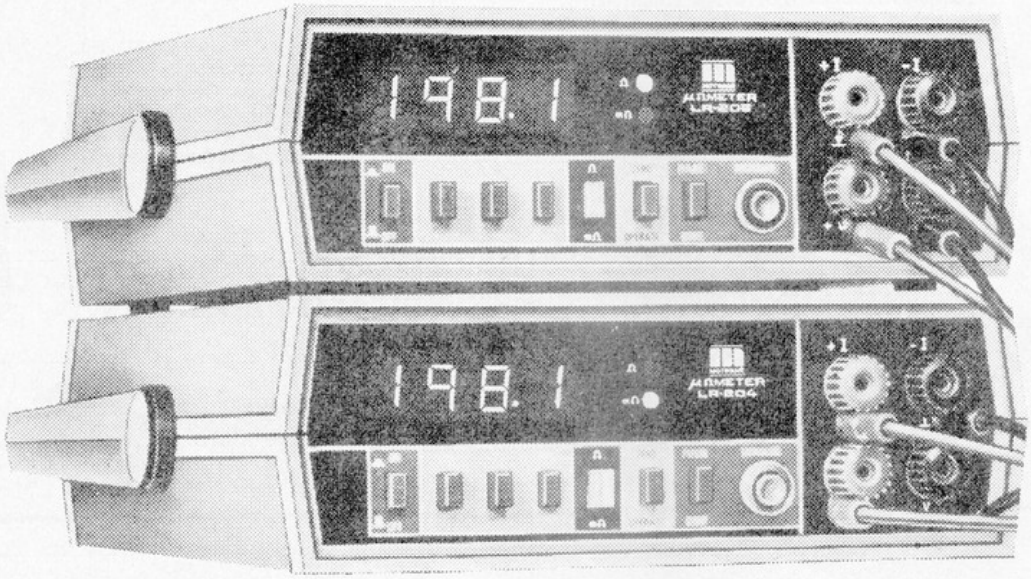

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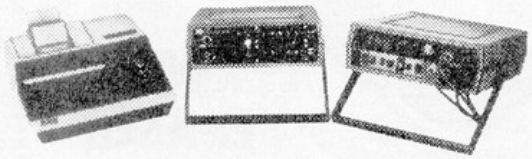

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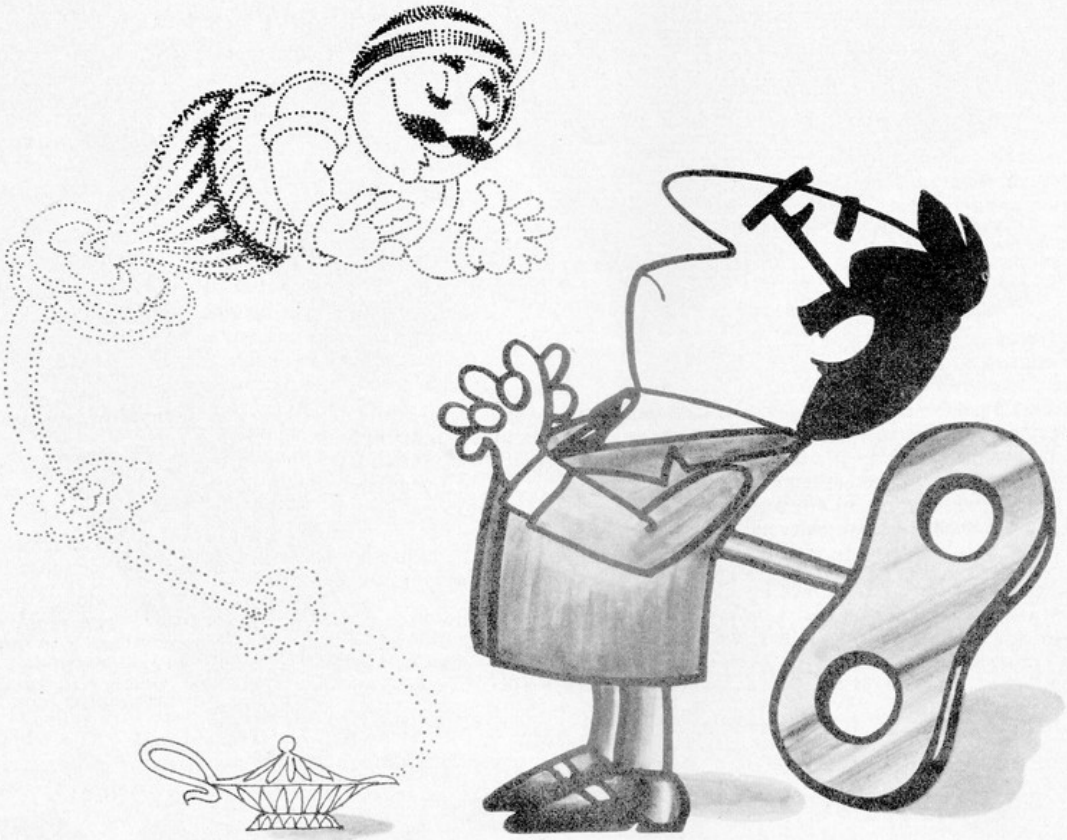

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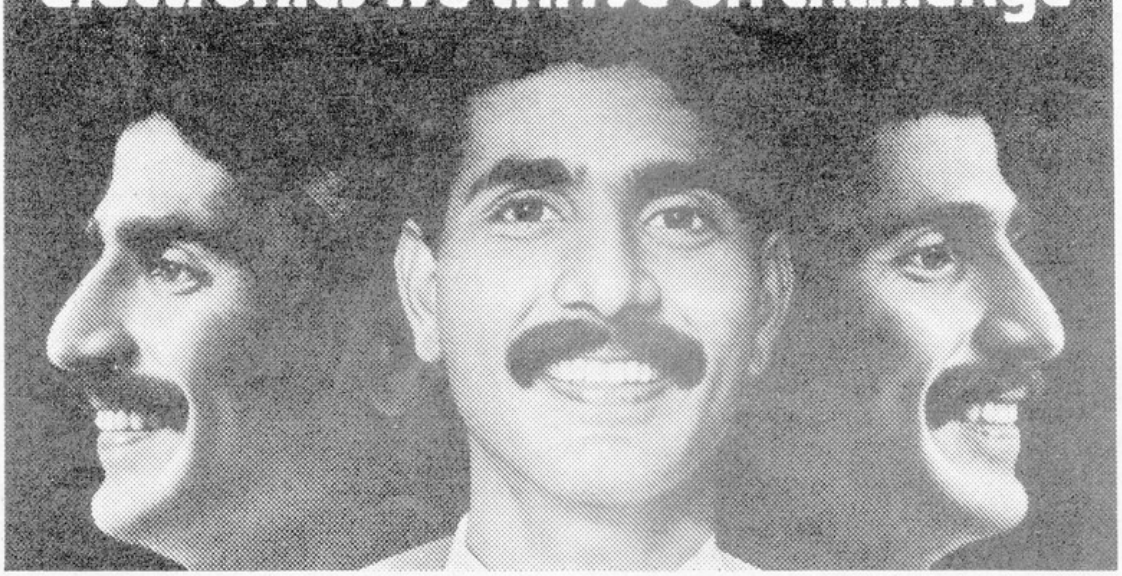


Multimedia Aquarius/AO/29/86

D/A CONVERTER FOR I/O BUS

by D Haehle

To further complete the range of plug-in I/O bus extensions, here is a programmable analogue output board. Those many owners of a C64, C128, or MSX micro equipped with the I/O bus board will, no doubt, appreciate the flexibility of the present extension, which is based upon the use of a readily obtainable D/A converter chip.

The C64 I/O bus, along with an associated digitizer board, was introduced in the June 1985 issue of *Elektor India*, while an 8-bit I/O port appeared in the December 1985 issue. How the universal I/O bus is to be modified to suit operation with the MSX series of computers is detailed in the *Elektor India*, February 1986.

The present D/A converter features 8-bit resolution, buffered outputs, and a presettable output voltage span. It is, we believe, an extremely simple to build module which will enable programmers to put their computer into contact with the real (analogue) world.

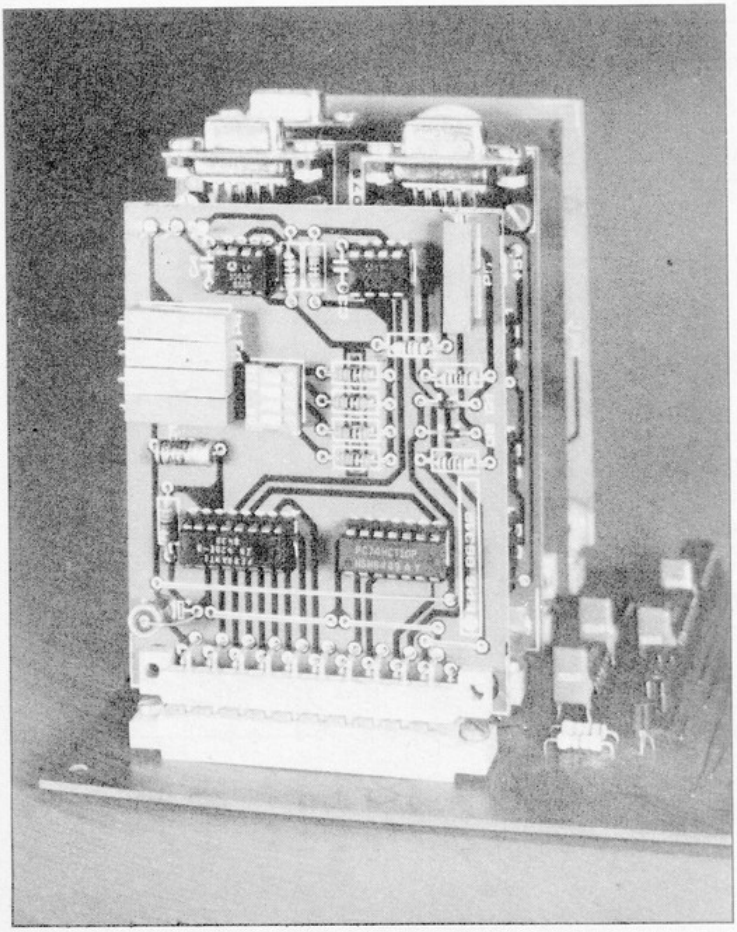
An 8-bit D/A converter chip

The proposed D/A board is based upon the use of the Type ZN428 8-bit digital-to-analogue converter (DAC), whose internal configuration is shown in Fig. 1.

A data latch loads the byte to be converted at the high-to-low transition of the latch pulse applied to the ENABLE input of the chip. The data will remain in the latch until a new byte is strobed into the device.

Each bit in the latch controls an associated electronic switch, S_1-S_8, connected to an R-2R resistor ladder network, which is fed from a stable reference voltage, U_{REF} (see Fig. 2). Depending on the magnitude (bit-configuration) of the latched byte, the switch poles of S_1-S_8 are either at digital ground potential, or at U_{REF}. Writing 255_{10} to the DAC latch, therefore, produces U_{REF} at the analogue output, since $255_{10} = FF_{hex} = 1111\ 1111$ (all switch poles at U_{REF}). Writing 0_{10} (00_{hex}), of course, produces an output voltage of nought, while 128_{10} (80_{hex}) yields $\frac{1}{2}U_{REF}$. In this manner, the chip output voltage can be stepped through the 0-U_{REF} range in 255 (2^8-1) increments.

Internal to the Type ZN428 is a 2.5 V reference voltage source, whose output is usually connected direct to the U_{REF} input of the R-2R array. It is also possible to use the U_{REF} output potential as a common reference for further D/A or A/D converters; in this way, system instability due to different temperature coefficients of individual boards can be ruled out quite effectively.



Circuit description

The circuit diagram of the D/A converter board is shown in Fig. 3. It is seen that the DAC latch contents are taken direct from the I/O board

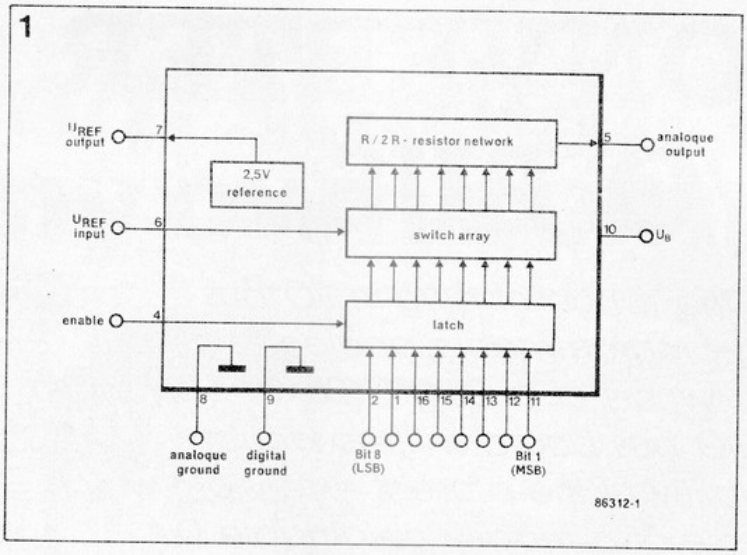


Fig. 1. Internal organization of the Type ZN428 8-bit digital-to-analogue converter.

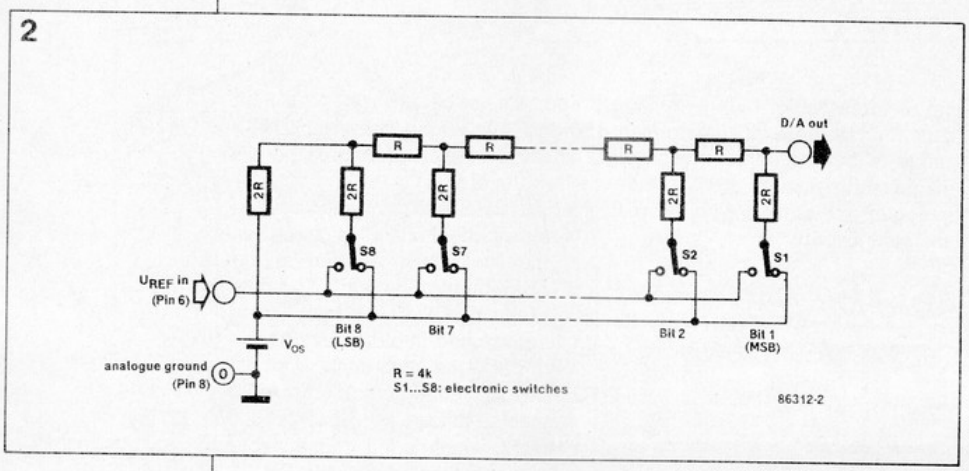


Fig. 2. Configuration of the R-2R ladder network internal to the ZN428 DAC.

databus, D_0-D_7. The latch enable pulse is obtained from gate network N_1-N_2-N_3 which has been set up for the logic function $\overline{EN}=(\Phi 2 \cdot \overline{SS} \cdot \overline{W})$. As already detailed in the article about the I/O bus, $\overline{SS}$ (Slot Select) is a signal which indicates the presence of a predetermined block of four addresses on the computer's address bus. Each slot (i.e. board on the I/O bus) is, therefore, accessible to the microprocessor via an appropriate memory read or write operation (PEEK/POKE). The present D/A converter board can be written to at any one of its four address locations.

The DAC reference is fed from the +5 V rail via R_1. C_1 has been added to effect the necessary decoupling.

The amplification of opamp A_1 can be set as required by means of DIL switches S_1-S_4. A_2 provides an output voltage that includes an offset level as defined with P_5.

The values of presets P_1-P_4 and associated series resistors R_2-R_5 are governed by the requisite amplification, A, of opamp A_1, according to

$A = 1 + R_a/R_b$ whence $R_b = R_a/(A-1)$.

With $R_a = R_6 = 10K$, and an object amplification of 4, R_b works out at 3K3. R_b is next divided into a fixed resistor and a preset, whence $R_b = R_2 + P_1$, and, in practice, $R_2 = 1K0$; $P_1 = 5K0$. These components ensure reaching the required amplification

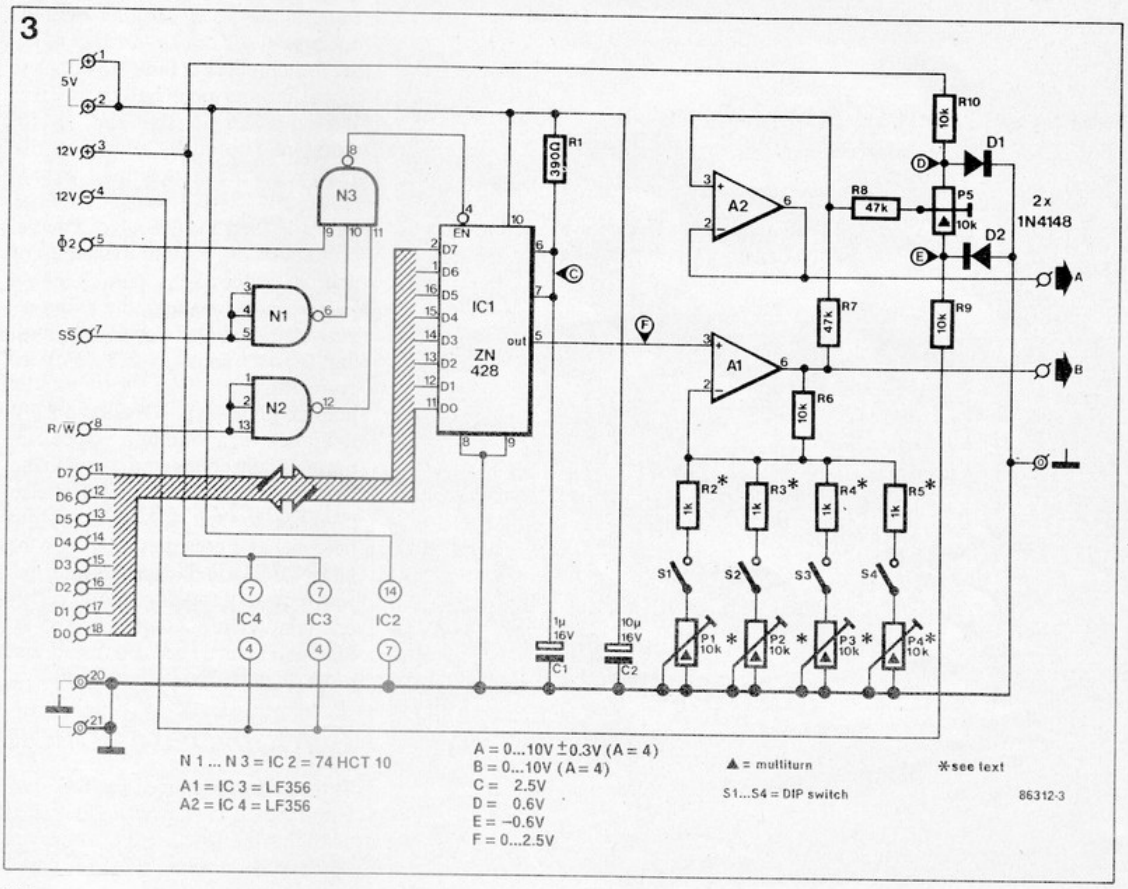


Fig. 3. Circuit diagram of the 8-bit DAC board. Its output voltage span and offset are user-definable.

factor (4) with P_1 set to roughly the centre of its travel. The dimensioning of the remaining resistor-preset combinations is, of course, identical to the above example with P_1-R_2.

The output voltage range at the A output is $0-(A \times U_{REF})$ V, while that at the B output is $V_{os}-(A \times U_{REF})+V_{os}$ V, where V_{os} is the offset voltage introduced with P_5.

Example: a VCO (voltage controlled oscillator) requires to be driven with 0.1-10 V. The amplification of A_1 must, therefore, be $10/U_{REF}=4$. With $P_1=5K0$ and $R_2=1K0$ (see the foregoing calculations), write 255_{10} to the relevant slot address and adjust P_1 for a DMM reading of 10.00 V at the B output. Write Ø to the DAC and adjust P_5 for a DMM reading of 100 mV at the A output.

Construction

By virtue of the simplicity of the present design, anyone with only limited experience in electronics construction should be able to get the board built and operative in a relatively short time. Fig. 4 shows the way in which the various parts are to be fitted onto ready-made PCB Type 86312. Do not forget the three wire links, and observe the correct orientation of the angled 21-way busconnector.

A demo program

Table 1 is a listing of a test and demonstration program intended to get the feel of programming the DAC in BASIC. It stands to reason that machine language routines can offer a considerable speed-up as compared with BASIC. For instance a programmable sine wave generator using the DAC board would require the user to be well acquainted with the intricacies of programming at the mnemonics level, which is, and should be, a real challenge in that it stands for mastery of both hardware and software.

St

4

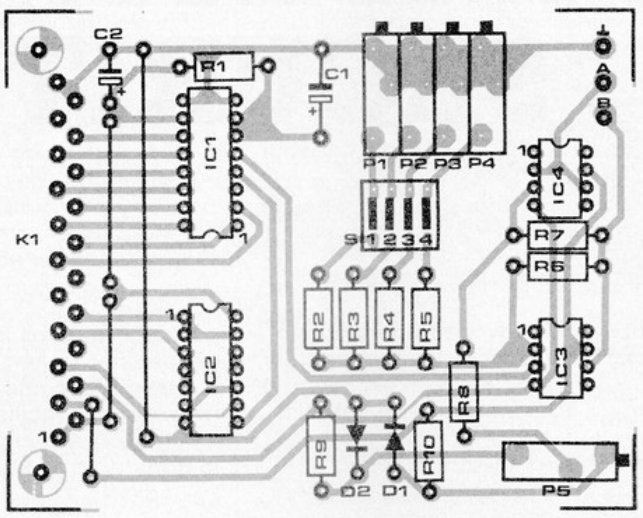


Fig. 4. Track layout and component mounting plan of the I/O bus DAC board. By virtue of its simple address decoder, the board may be plugged into any slot on the I/O bus-board.

Parts list

(Note: parts are coded to BS 1852; see Infocard 500)

Resistors:

(tolerance is ±5% unless otherwise stated)

$R_1=390R$

$R_2 \ldots R_5$ incl. = 1K0 *

$R_6;R_9;R_{10}=10K$

$R_7;R_8=47K$

$P_1 \ldots P_4$ incl. = 10K multiturn preset *

$P_5=10K$ multiturn preset

Capacitors:

$C_1=1\mu0$; 16 V tantalum

$C_2=10\mu$; 16 V tantalum

Semiconductors:

$D_1;D_2=1N4148$

$IC_1=ZN428$ (Ferranti)

$IC_2=74HC(T)10$ or 74LS10

$IC_3;IC_4=LF356$

Miscellaneous:

$S_1 \ldots S_4$ incl. = 4-way DIL switch

$K_1=21$-way angled busconnector to DIN41617

PCB Type 86312 (see Readers Services)

3 off soldering pins

* See text for dimensioning details.

Table 1

```
 10 REM  ********************************
 20 REM * TEST PROGRAM FOR I/O BUS       *
 30 REM * SLOT LATCH AT $E120 = 57630    *
 40 REM * SLOT 1 = $E120 - E123 = 57630 - 57633  *
 50 REM * SLOT 2 = $E124 - E127 = 57634 - 57637  *
 60 REM * SLOT 3 = $E128 - E12B = 57638 - 57641  *
 70 REM * SLOT 4 = $E12C - E12F = 57642 - 57645  *
 80 REM  ********************************
 90 REM
100 REM ********** INPUT DATA **********
105 PRINT:PRINT
110 INPUT" WHICH SLOT HOLDS 8-BIT DAC BOARD (1-4)";S
115 ON S GOSUB 1000, 1010, 1020, 1030
120 INPUT" AMPLIFICATION FACTOR              ";A
130 INPUT" OFFSET VOLTAGE        (VOLTS)     ";O
140 INPUT" REFERENCE VOLTAGE     (VOLTS)     ";RV
150 INPUT" OUTPUT VOLTAGE AT A   (VOLTS)     ";VO
160 REM
300 REM ********** COMPUTE LATCH CONTENTS *****
310 REM
320 X1=(VO-O)/A
330 X2=(X1/2.56)*RV*100
340 IF X2>255 THEN X2=255
350 POKE B,X2
360 REM
400 REM ********** DISPLAY VALUE **********
410 REM
420 PRINT:PRINT" VOLTAGE = ";VO;" VOLT":PRINT:PRINT
430 FOR I=1 TO 1000:NEXT I
440 REM
500 REM ********** LOOP **********
510 REM
520 GOTO 150
530 REM
1000 B=57632:RETURN
1010 B=57636:RETURN
1020 B=57640:RETURN
1030 B=57644:RETURN
2000 END
```

Table 1. Simple BASIC program to verify the correct operation of the 8-bit DAC board. Observe the I/O map addresses at lines 1000-1030, and, if necessary, modify these to suit your computer's I/O address configuration.

COMMUNICATIONS SECURITY

The problem of unauthorized access to information in transit is as old as mankind itself. Each advance in the means of transferring information has tended to be accompanied by more sophisticated possibilities for compromising that information. Thus, in 1986, despite the millions of possible frequency hopping sequences it may possess, the integrity of a modern military radio which falls into enemy hands is assumed to be no more than a few hours. However, there has been a fundamental change in one aspect of this sector in recent years. Traditionally, interest in secure communications was confined to the military and diplomatic communities.

Today, because of the increased dependence on the continual exchange of information—be it technical, financial, or personal—the problem applies equally to industry, commerce, and public administration.
Consequently, an industry dedicated to interfering with communication terminals of all sorts has evolved. Products available range from local and remote telephone bugs, through microtransmitters on terminal keyboards, to systems that can reconstruct conversations from electromagnetic emissions from terminals.
Illicit access to information takes many forms, from the accidental interception of a conversation (as in the case of a crossed telephone line), to the planned electromagnetic attack on a specific communication system known to carry data of a sensitive nature.

Disguising messages

In the 20th century, the consequences of compromising privileged information vary from personal embarrassment to considerable pecuniary loss. Illicit access to personal information—for example, credit rating—can have serious cumulative effects for the individual concerned. Unauthorized access to other information—perhaps that relating to impending changes on the stock exchange, proprietary processes, or movement of high value cargoes—can result in obvious financial loss.
Malicious modification of information in transit can also have dramatic results.
The science of avoiding such consequences can be divided into two parts—the protection of the media and the defence of the message.
In practice, the comprehensive protection of modern communications media is not a practical proposition. Metal communication cables, to an extent, act as radiating antennas and the confidentiality of traffic can be jeopardised by the use of suitable listening equip-

The Plessey Crypto Voicelok 100 secure telephone provides complete security for the executive who has to discuss sensitive matters on the telephone.

ment. The most vulnerable types are the overhead open wire links that are common in some parts of the world. Screened and shielded cables, particularly if buried, are a more difficult proposition for the would-be eavesdropper, but physical access to the cable makes interception relatively straightforward. Fibre optic cable, which carries photon instead of electrical energy, does not act as a radiating aerial. Attempts to tap into fibre should shut the whole link down, although it is believed that considerable sums of money are being invested in systems to surmount this impediment to eavesdropping. Radio signals can also be compromised by sensitive receiving systems. As a generalization, the more directional and precise a transmission, the more difficult is the process of compromising it. Tropospheric scatter and meteor burst communications score highly for some specialized traffic types.

These interception risks have focused increasing attention to disguising the content of messages.

Voice scrambling involves the partitioning, rearrangement, and permutation of the signal, with the reverse processes being applied at the receiving end. An interesting development in this sector is the FX204 scrambler from Consumer Microcircuits[1], a new monolithic implementation for radio application including cellular and cordless telephones.

Digital vs analogue

In essence, the FX204 is a two-band frequency inversion device that uses switched capacitor filters to split the voice spectrum into high and low frequency bands, and balanced modulators to invert each frequency band about its own centre frequency.

The split point frequency is externally programmable to 32 different points in the range 300 to 3000 Hz and makes the FX204 suitable for both fixed programmable and rolling code speech scramblers. All filter cut-off frequencies and inversion carriers are derived from a single reference crystal oscillator and facilities are provided to input and output synchronization tones where required. Constructed in a 5 V single supply, complementary metal-oxide semiconductor (CMOS) process and available in dual-in-line (DIL) and surface-mounted 24-pin packages, the FX204 is suitable for use in fixed or portable equipment.

Encryption schemes, which are usually digital, are applied to both voice and data signals. They operate by reducing the signal to a bit stream which is then permutated and transposed on a bit-by-bit basis according to the dictates of a key which is applied at the receiving end.

A comparison of analogue and digital systems reveals that the former provide a more natural voice quality and allow speaker recognition. Digital systems, by synthesizing the original signal, are not as competent in this area. In terms of traffic security, it is recognized that digital systems in general operate at higher levels, although some analogue arrangements are comparable. Some digital systems require data compression and so are more expensive for voice than analogue systems which can transmit over conventional speech channels without modification. Analogue signals become dirtier than digital ones during the course of amplification on long distance metal carriers.

A critical difference between the two lies in the likelihood that an analogue system will be vulnerable to the engineer and the cryptoanalyst, the sophistication of their equipment, and the complexity of the original scrambling technique. Digital systems are more readily dealt with by computer facilities; here the concerns are the key algorithm and the signal.

Protection standards

Advances in technology have made it possible to build complex and sophisticated encryption equipment for use over the public switched telephone network. Plessey Crypto[2], for example, has recently introduced the Voicelok 100 secure telephone. Voicelok, which uses a patented, essentially analogue time division technique, has 10^{12} possible key settings. It is designed for high-level security over telephone networks and is available in either multi-frequency or loop-disconnect signalling versions. All encryption circuits are contained on a single module within the telephone. Based on the Plessey PBT 100 series, Voicelok has two operating modes—clear and secure. In the clear mode the instrument is a standard duplex telephone link. Switching to secure mode by an illuminated push button initiates a 16 bit key, pseudo randomly generated, and transmitted at each transmission path reversal. This involves a transmission delay of typically 600 ms. Plessey has also developed Faxlok, a system using similar technology, for facsimile transmission.

Widely used internationally is the Data Encryption Standard (DES) developed in the United States of America by IBM. DES uses keys that are periodically changed, either physically or electronically. Due to the American government's restrictions on the export of high technology, there have been periodic shortages of encryption devices embodying the DES algorithm. It has also been reported that the American National Security Agency (NSA) does not intend to recertify DES when it is reviewed in 1988.

These developments have led organizations throughout the world to develop alternative encryption standards. A new encryption chip, developed by British Telecom[3] and called B-crypt, is a device that embodies such a standard. BT says that in some respects its B-crypt devices are superior to DES devices. In particular, data sent over telephone lines contain a lot of repetitive elements such as data address messages or headers. DES encrypts these in the same way as the main data, thereby giving clues to the cryptoanalyst, while B-crypt encodes headers in a different way.

BT is also working on a telecommunications authorities cryptographic algorithm (TACA), designed to protect data sent over satellite circuits.

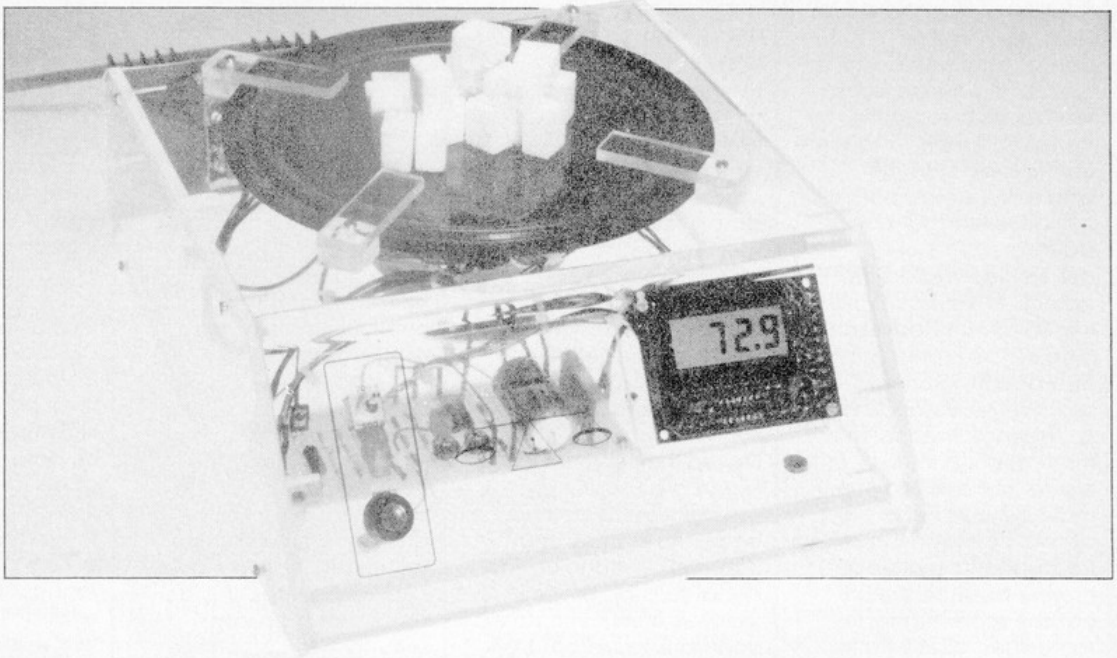
(1) Consumer Microcircuits Ltd; Wheaton Road; Industrial Estate East; Witham; Essex CM8 3TD.

(2) Plessey Crypto; Wavertree Boulevard; Wavertree Technology Park; Liverpool; Merseyside L7 9PE.

(3) British Telecom Centre; 81 Newgate Street; London, EC1A 7AJ.

ELECTRONIC BALANCE

by R Ochs



This accurately operating balance, which is entirely composed of electronic parts, features a 3½ digit read-out, a tare offset facility, and a weighing capacity of 500 grammes. Based upon the use of a common bass loudspeaker as the weight sensor, this novel household utensil is readily built and extremely useful for a variety of hobby applications, and, of course, for cooking!

Like most types of electronic balance, the proposed low-cost version is based on the underlying principle of electromagnetic force compensation. Since the force on a conductor placed in a magnetic field is proportional to the coil current causing the field, the voice coil in a loudspeaker can be used as a force sensor, if weight is transferred direct onto the cone and thus onto the voice coil. After measuring the cone displacement, an electronic control circuit arranges for a current to be sent through the voice coil, causing the initial position of the cone to be shifted, i.e. it is pushed outwards. The current necessary to effect the counterbalancing cone displacement is directly proportional to the force applied to the voice coil. In the proposed design, the loudspeaker is a fairly powerful type with a flexible cone suspension system that ensures adequate repeatability in the stated weight ranges of 0 to 200 and 200 to 500 g. Also, the loudspeaker should be capable of handling considerable dissipation, as its voice coil is fed with a direct, rather than an alternating (AF), voltage. The foregoing considerations regarding the requisite type of loudspeaker leave virtually no other choice than a rugged woofer with a power handling capability of some 100 W.

The weight sensor

Converting the loudspeaker into an accurate weight sensor is not too difficult, provided the cone, membrane and voice coil are treated with care. A pre-heated knife may be used to loosen and remove the dust cover in the cone centre. Once you have gained access to the magnet and voice coil assembly, great care must be taken to prevent small metal parts or even dust from entering the air gap, since this will have a highly adverse effect on the linearity of the balance.

Fig. 1 illustrates how to proceed with the construction. The light barrier is carefully glued onto the magnet, and its three wires to the control circuit are left long enough to allow for the maximum anticipated cone displacement, before they are fed through small holes in the cone, glued into place, and connected to a terminal strip fitted onto the loudspeaker chassis.

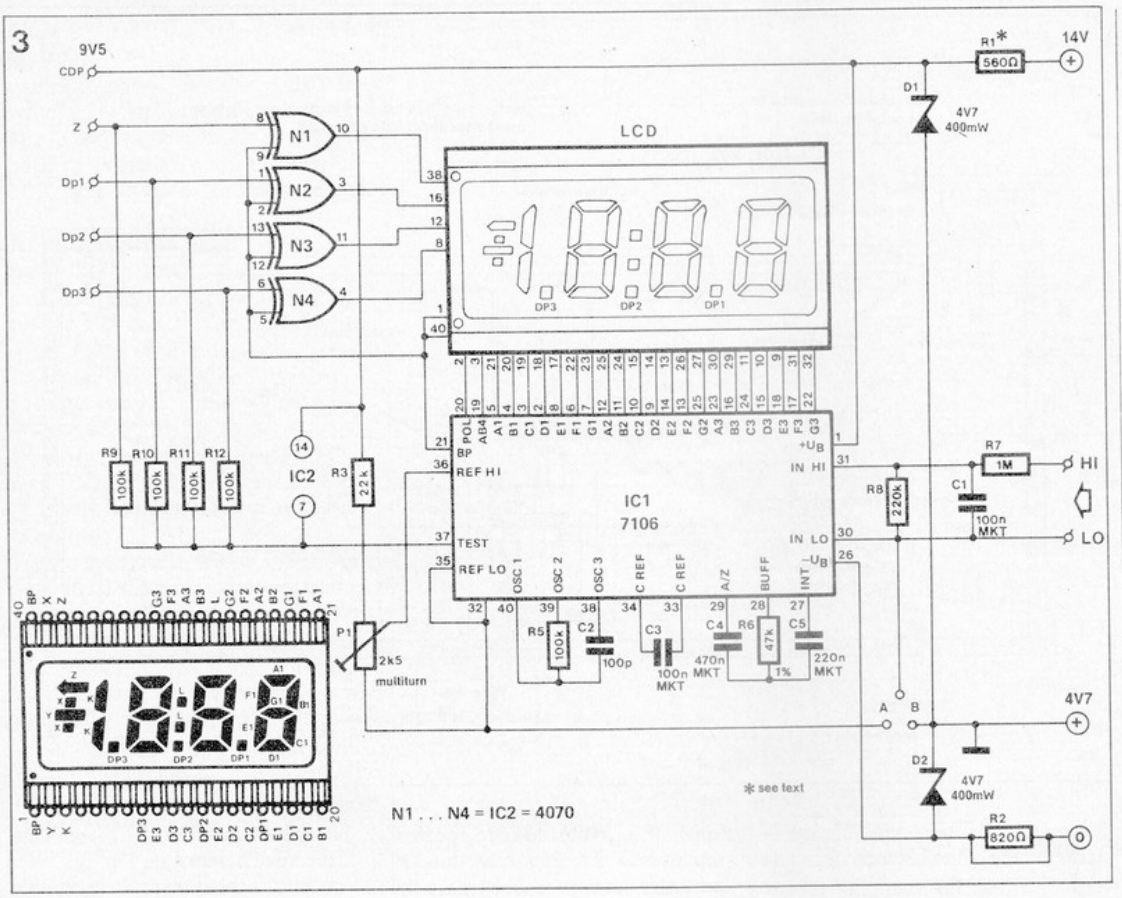


Fig. 3. Circuit diagram of the 7106-based digital readout.

Parts list (see Fig. 4)

Resistors:
$R_1 = 560\ \Omega$
R_2 = wire link
$R_3 = 22$ k
R_4 = not required
$R_5; R_9 \ldots R_{12}$ incl. = 100 k
$R_6 = 47$ k
$R_7 = 1$ M
$R_8 = 220$ k
$P_1 = 2k5$ preset

Capacitors:
$C_1; C_3 = 100$ n
$C_2 = 100$ p
$C_4 = 470$ n
$C_5 = 220$ n

Semiconductors:
$D_1; D_2 = 4V7; 0.4$ W zenerdiode
$IC_1 = 7106$
$IC_2 = 4070$

Miscellaneous:
LCD = 3½ digit liquid crystal display; digit height 13.3 mm (e.g. Hamlin Type 3901 or 3902 SE 6902)
PCB Type 84012-2 (see Readers Services)
Suitable sloping front cabinet

Most likely, you will find that there is a tendency to oscillate at relatively low weights with P_2 adjusted for a high P/I ratio, while increasing the integrated (I) portion promotes oscillation at relatively large weights. If attempts to stabilize the cone movement are unsuccessful, the system may have to be pre-loaded with a small weight; however, this should not normally be necessary if the loudspeaker is adequately damped by the enclosure lining.

The adjustment screw (see Fig. 1) should be set to produce a slight upward cone displacement at power-on; the voice coil quiescent current should then lie between 10 mA and 50 mA.

Both weight ranges are calibrated by adjusting the 7106 gain preset, P_1 (see Figs. 3 and 4) for an LCD read-out that tallies with a few standard weights placed onto the platform. Alternatively, but with some loss in accuracy, a number of small weights may be made at home by wrapping sugar lumps into paper and having these weighed at a chemist's.

TW

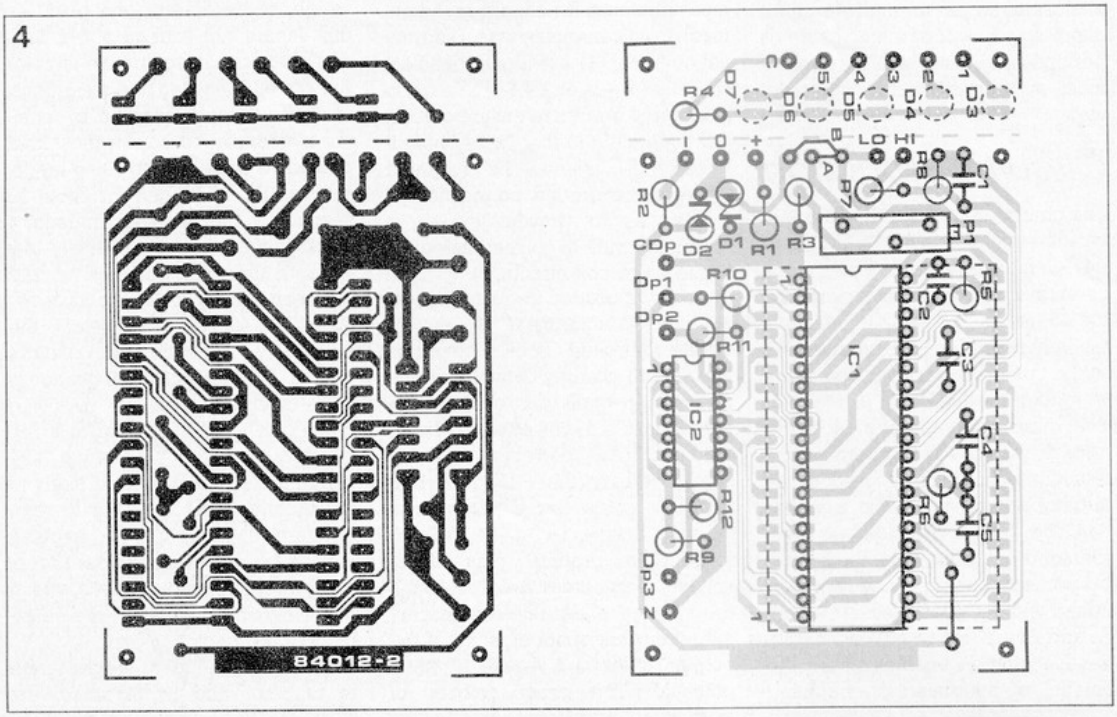
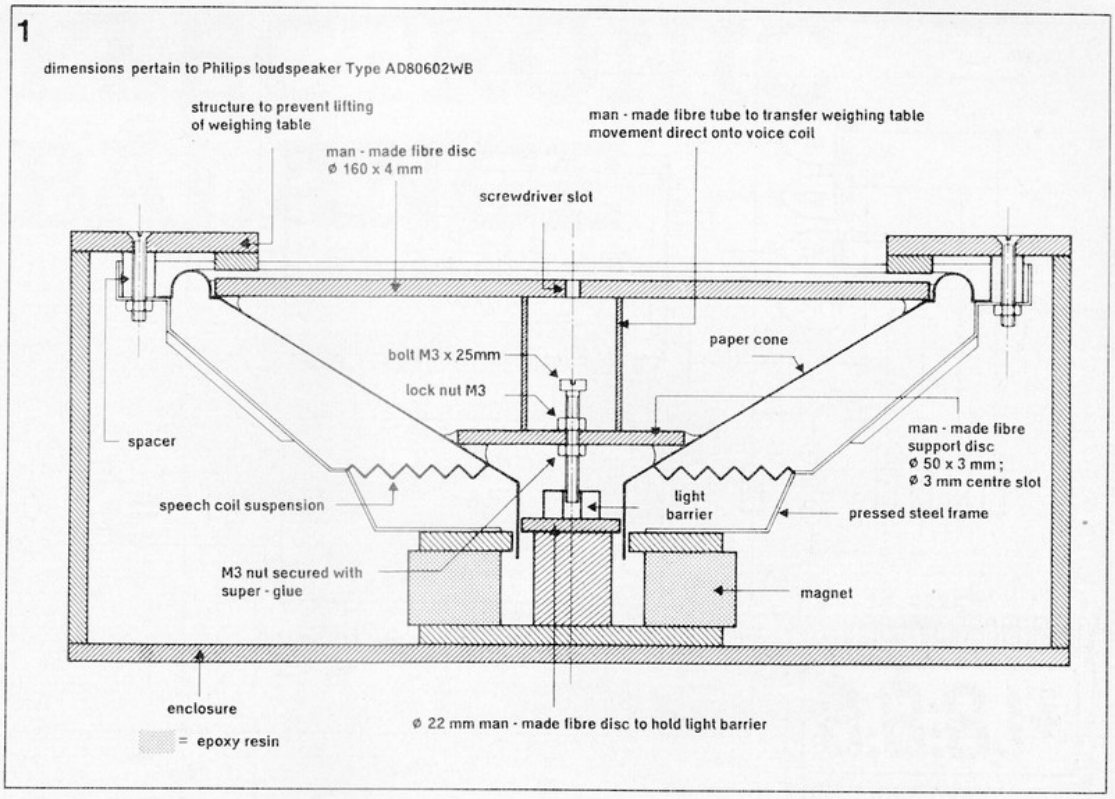


Fig. 4. Track layout and component mounting plan of the LCD board in the electronic balance. Fit neither link A nor link B.

Fig. 1. Cross-sectional view of the balance enclosure.



It should be borne in mind that the final accuracy of the balance depends very much on the damping of the loudspeaker; since the control circuit is a proportional integration type (PI, this will be reverted to), the removal of fairly heavy weights from the platform may be counteracted rather slowly, causing forceful cone displacement and oscillation at very low frequencies. Therefore it is strongly suggested to fit the loudspeaker and associated control & supply circuits into an air-tight enclosure so as to improve upon damping. A wooden enclosure is perfectly adequate, both from a technical and an aesthetical point of view.

Circuit description

The control circuit in the electronic balance is shown in Fig. 2. Light barrier IC_3 functions as the sensor, since its output voltage is determined by the adjustment screw that interrupts the light beam from the internal LED as the cone sinks deeper due to the weight on the platform.

The current-control loop is based upon the use of a PI (proportional integration) circuit, composed of integrator A_2 and adjustable amplifier A_3. The former provides a time-averaged output voltage, the latter a proportional output voltage determined with feedback preset P_1. Both A_2 and A_3 are driven by input amplifier A_1, while P_2 enables setting the amount of integrated or amplified signal (P/I ratio) fed to current amplifiers T_1-T_2. Potentiometers P_1 and P_2 are set to positions where the control loop output signal is free from oscillation. As stated above, fitting the balance in a closed cabinet is the best way to go round this problem.

Current sense resistor R_{19} drops a voltage in direct proportion to the current passed through the voice coil. In order to achieve a relatively low temperature-coefficient and hence optimum repeatability of measurements, R_{19} should be home-made from constantan wire. Differential amplifier A_4 has a gain of 20 dB. Note that R_{13} is at +4.7 V relative to the supply ground to ensure correct DC interfacing to the display board. Resistor R_{10} should be mounted close to the ground connection of R_{19} so as to prevent erroneous readings owing to contact resistance in the voice coil circuit.

The circuit around A_5 is a sample-and-hold arrangement to enable switch-controlled tare subtraction. Pressing S_2 charges C_7 with the output voltage of A_4 and at the same time forces a reset of the balance read-out. At power-on, C_7 is discharged and the + input of A_5 is therefore at the same potential as junction R_{14}-R_{15}, i.e. at +4.7 V with respect to ground, plus 80 mV voltage drop across R_{15}. The 80 mV drop serves to establish a quiescent output current of about 40 mA if the outputs of A_4 and A_5 are at equal potential. The exact amount of quiescent current can be set with the adjustment screw (see Fig. 1).

The tare/reset button, S_2, is simply pressed after determining the weight of the tray, jar, or any other container which is to hold the relevant substance for weighing. In a similar fashion, S_2 can be used to reset the display prior to adding a further ingredient to a mixture, according to the recipe to hand. There are, however, a few important points to observe in the use of the tare facility. The first concerns the total weight of the load on the platform: this should not exceed 500 g. Second, there is a specific time limit for pressing S_2 between tare weighings, as C_7 is slowly discharged by its internal resistance and the load presented by A_5. In the 200 g range, tare weight is retained for about 30 seconds, in the 200-500 g range for a much longer time. At relatively low weights, therefore, readings should be taken rapidly for best accuracy.

Switch S_3 is used to select the previously mentioned weight ranges. Although the 0-200 g range is more accurate than the 200-500 g range, the former calls for S_2 to be pressed prior to any weighing. The pre-set quiescent current is likely to be slightly instable owing to temperature changes in the cabinet, caused by voice coil, current loop, and power supply dissipation, which has a negative effect on the sensitivity of the phototransistor.

Selection of the higher weight range is accomplished by S_3 taking the

voltage from divider network R_{16}-R_{17}-R_{18}. At the same time, the decimal point on the LCD is switched to the appropriate position.
The digital read-out for the proposed balance is based upon the use of PCB Type 84012-2, incorporated in the Capacitance Meter, published in the March 1984 issue of *Elektor India*. The circuit diagram is shown in Fig. 3, the component mounting plan in Fig. 4. Note that *neither link A nor link B* should be fitted on the board to suit operation with the balance control circuit. Also note that D_1-D_2-R_1 are in fact the parts shown to belong to the power supply (see Fig. 2); they are most conveniently fitted onto the LCD board, of course.
Functional details of the Type 7106 LCD display driver have been discussed in *LCD panel meter, Elektor Electronics* October 1981. In brief, and with reference to Fig. 3, R_5 and C_2 determine the internal oscillator frequency of about 45 kHz, which is used to derive the sample and measurement interval. Capacitor C_4 functions as the auto-zero capacitance, which, correctly dimensioned, ensures a reading of 000 on the LCD with both chip inputs at 4.7 V. The maximum indication on the LCD is reached at an input voltage of $2V_{REF\ HI}$; therefore P_1 determines the final sensitivity of the display board.
The power supply for the proposed balance can source up to 1.5 A and requires adequate cooling. The stated value of R_3 gives an output voltage of 14.0 V, which determines to a large extent the voice coil current at maximum weight, i.e. 500 g. Network R_1-D_1-D_2 serves to stabilize the LCD board supply voltage, and to create a virtual common rail at +4.7 V above the circuit ground potential. The +9.5 V potential for sample-hold A_5 is taken from point CDP on the LCD board. The voltage limit prevents the 7106 inputs from being driven in excess of their handling capability.

Setting up

To begin with, the PI control loop should be adjusted. This may be a little tricky in view of the previously mentioned tendency to oscillation at low frequencies. Moreover, oscillation may occur with different weights on the platform.
Checking for undesirable oscillation at low frequencies is best done by monitoring the output of A_1 with a DC-coupled oscilloscope, while the weight is increased slowly by piling sugar lumps on the weighing table.

2

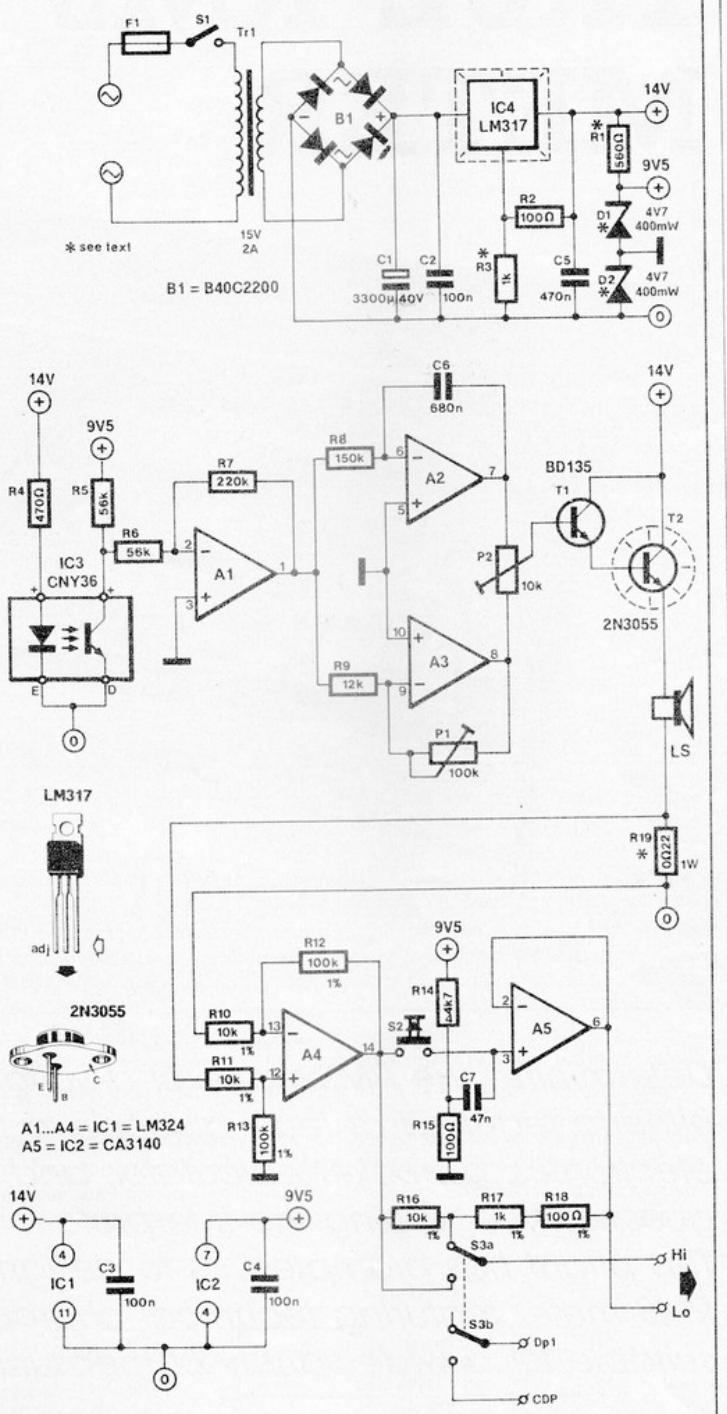


Fig. 2. Circuit diagram of the balance control and supply sections.

Parts list (see Fig. 2)

Resistors:
R_1 = 560 Ω▼
R_2;R_{15} = 100 Ω
R_3 = 1 k
R_4 = 470 Ω
R_5;R_6 = 56 k
R_7 = 220 k
R_8 = 150 k
R_9 = 12 k
R_{10};R_{11};R_{16} = 10 k;1%
R_{12};R_{13} = 100 k;1%
R_{14} = 4k7
R_{17} = 1 k;1%
R_{18} = 100 Ω;1%
R_{19} = 0Ω22 *
P_1 = 100 k
P_2 = 10 k

Capacitors:
C_1 = 3300 µ;40 V
C_2...C_4 incl. = 100 n
C_5 = 470 n
C_6 = 680 n
C_7 = 47 n

Semiconductors:
D_1;D_2 = 4V7;0.4 W ▼
T_1 = BD135
T_2 = 2N3055
IC_1 = LM324
IC_2 = CA3140
IC_3 = CNY36
IC_4 = LM317T

Miscellaneous:
S_1 = SPST mains switch
S_2 = push to make button
S_3 = miniature DPDT switch
F_1 = 200 mA delayed action
LS = bass loudspeaker (woofer); 8 Ω; ϕ 200 mm; P_{min} = 60 W (e.g. Philips Type AD80602W8) or Rencforce 60/100 W 8 Ω).
B_1 = bridge rectifier; 2 A min.; e.g. B40C2200
Tr = mains transformer 15 V;2 A.

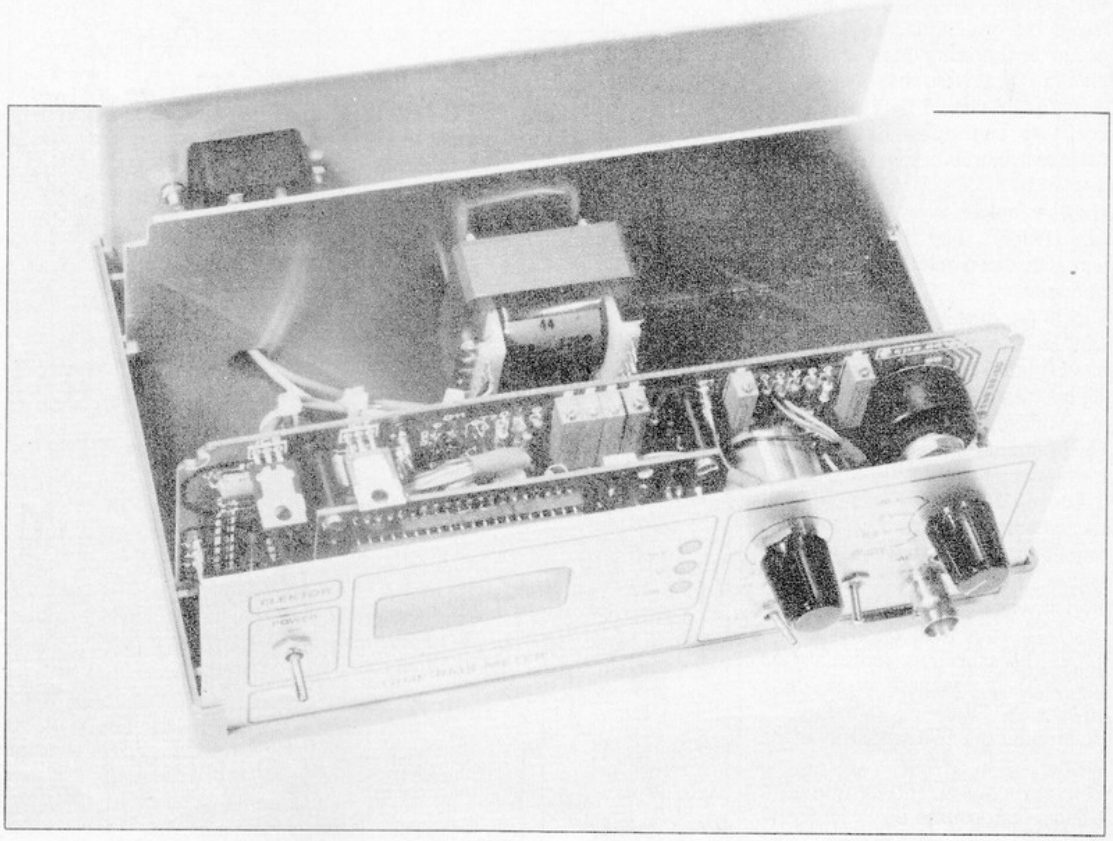
▼ On display board
* See text

We regret that no ready-made PCB is available for this part of the circuit.

Technical characteristics:

Weight ranges:	0.....200 g and 200...500 g.
Maximum weight load:	500 g.
Linearity:	<1% of read-out ±1 digit.
Accuracy:	<0.5% of full-scale indication ±1 digit (= 0.1 g in 200 g range).
Compensation for off-centre placed weights:	<2% of read-out at weighing table diameter of 100 mm.
Loudspeaker:	ϕ 200 mm; 60-100 W, 8 ohm.
Display:	3½ digit, switched decimal point.

TRUE RMS METER



Determining the RMS value of a voltage or current hitherto required at least a scope, a textbook of basic electronics, a pocket calculator, and, at times, sheer guesswork regarding the interpretation of resultant figures. This plight has prompted us to design a wideband AF RMS meter featuring technical characteristics to make it suitable for a wide variety of measuring applications.

Technical characteristics:

Input ranges:	20 mV; 0.2 V; 2 V; 20 V (−40 dB; −20 dB; 0 dB; +20 dB).
Accuracy ($U_{in} = \frac{1}{2} U_{in(max)}$):	±(1.5% + 1 digit) 0-100 kHz; ±5% 100-200 kHz.
Bandwidth ($U_{in} = \frac{1}{2} U_{in(max)}$):	300 kHz (−3 dB).
Span of variable 0 dB level:	+65 to −30 dB.
Special features:	switch-selectable 0 dB reference; AC or DC coupled input (20 mV AC only); 3½ digit LC display; optional LIN and LOG outputs to drive external instruments.

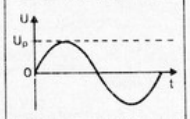
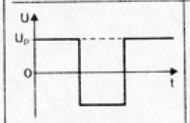
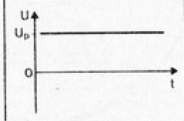
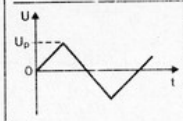
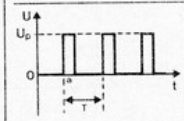
In electronics literature, both at the hobby and the professional level, the synonyms *effective*, *virtual*, and *root-mean-square* are frequently used to qualify an alternating quantity such as voltage or current. Also component ratings, maximum permissible dissipation, AF and RF signal levels, to name but a few examples, are frequently stated as being rms values. For relatively low-frequency, pure, sinewaves, the rms amplitude can be read with ample accuracy from an analogue or digital AC voltmeter, since these instruments are generally calibrated for the sine wave *crest*

factor of $\sqrt{2}$. However determining the rms value of other periodical signals, such as ramp voltages, rectangular, or triangular waveforms, is not usually possible with the same AC voltmeter for lack of a calibration in accordance with the requisite crest factor, defined as the ratio of the peak value of a periodically varying quantity to the root-mean-square value. Without going into mathematical details, Table 1 summarizes terms and conversion formulas for some of the most frequently encountered waveforms (see also Infocard 114).

The proposed meter is based upon the rms-to-DC conversion principle and fulfills a variety of applications as it has been designed to accept many waveforms at wide frequency and input voltage ranges, ensuring a high input impedance. The instrument combines the functions of true-RMS meter and dB (AF input voltage level) meter, offering instantaneous readings on a liquid crystal display.

As seen from the introductory photograph to this article, the RMS meter adds to the Elektor range of measuring equipment housed in a standard Verobox enclosure. Extremely straightforward to operate and fairly easy to construct, the latest addition to the series achieves a remarkable degree of precision at moderate cost.

Table 1.

Waveform	effective voltage U_{rms}	average voltage U_{av}	form factor U_{rms}/U_{av}	crest factor U_p/U_{rms}
	$U_p/\sqrt{2}$ $\approx 0.707U_p$	$2U_p/\pi$ $\approx 0.637U_p$	≈ 1.111	$\sqrt{2} \approx 1.414$
	U_p	U_p	1	1
	U_p	U_p	1	1
	$U_p/\sqrt{3}$	$\frac{1}{2}U_p$	$2/\sqrt{3} \approx 1.156$	$\sqrt{3} \approx 1.732$
	$U_p\sqrt{a/T}$	$U_p(a/T)$	$1/\sqrt{a/T}$	$1/\sqrt{a/T}$

Block diagram

Fig. 1 shows the functional organization of the true-RMS meter. The AC or DC-coupled input voltage is fed to an amplifier/attenuator circuit, which ensures a maximum input level of 200 mV for the rms-to-DC converter. This means that the input section functions as an amplifier in the 20 mV (AC only) and 200 mV ranges (A = 10× and A = 1×, respectively), while it functions as an amplifier in the 2 V and 20 V input ranges (A = −10× and A = −100×, respectively). Selection of the relevant range is accomplished with an electronic switch arrangement, which obviates the drawbacks associated with long wires at relatively high impedance.

The rms-to-DC converter provides a linear as well as a logarithmic direct output voltage. With S_5 set to the V position, the linear output voltage is fed direct to the analogue-to-digital converter comprised in the LCD read-out circuit. With S_5 set to the dB position, the display unit accepts the HI input voltage from a temperature compensation circuit connected to the converter's logarithmic output. This compensation circuit is based upon the use of an amplifier whose gain is temperature-dependent and whose output is applied to a voltage divider to achieve a 1 mV/dB gradient with respect to ground.

Provision has been made to select either a fixed or a variable (offset) 0 dB threshold (0 dB = 0.775 V ≙ 1 mW into 600 ohms, see also Infocard 501). The taking of linear rms readings is quite straightforward in that it merely involves selecting the

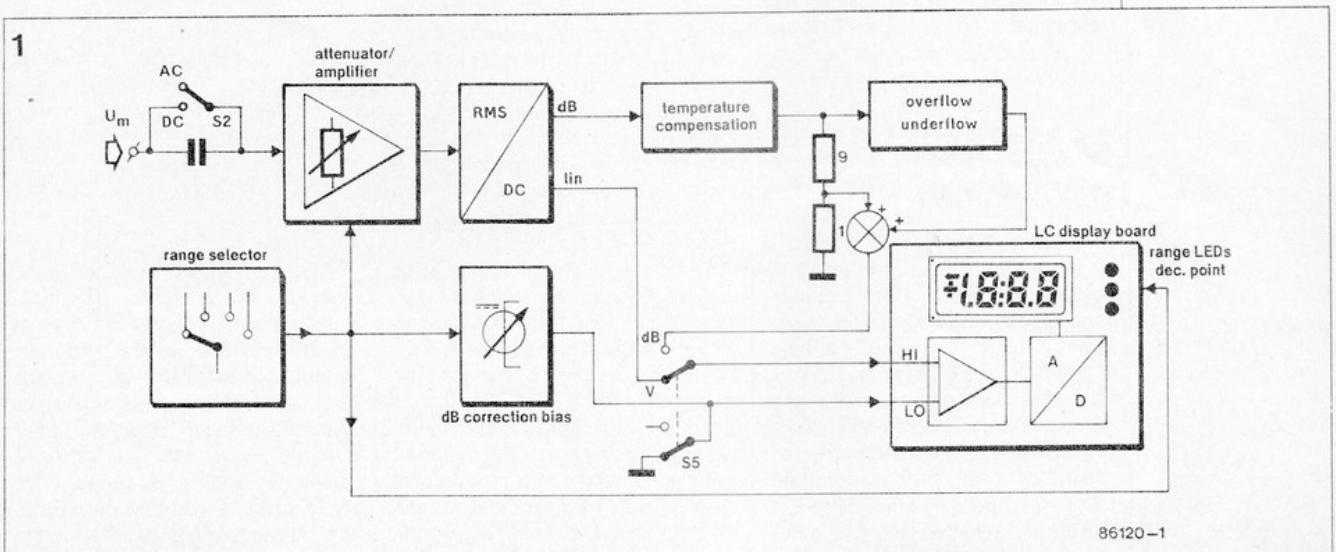


Fig. 1. Block diagram of the true-RMS voltmeter.

appropriate attenuation or amplification factor of the input section, plus switching the decimal point on the LC display. There is a snag, however, in the reading of dB levels. Assuming a meter input level of 0 dB (0.775 V_{rms}), the rms converter chip is fed with 77.5 mV (input attenuation is 10 times in the 2 V/0 dB range) and it can be adjusted to yield the correct LCD reading. However should the meter be switched to its +20 dB input range, the input voltage is attenuated 100 times, and the converter input voltage is, therefore, 7.75 mV, which would cause the display to read $20\log_{10}(7.75/77.5) = -20$ dB, rather than still 0 dB. This error is corrected by applying −20 mV to the LO input of the LC driver. A similar correction applies to the −20 dB and −40 dB ranges, in which case LO is driven with +20 mV and +40 mV, respectively.

An overflow/underflow circuit provides users of the meter with information as to the preferred range for use with a specific input voltage level. Should this exceed the maximum displayable value by about 14 dB, the LC display gives an overflow indication. Similarly, an input level of 30 dB below the set value is signalled with the underflow sign, prompting the user to switch to the next lower range for optimum accuracy. However in the −40 dB range of the meter the underflow circuit is disabled to allow carrying out measurements at very low input levels. It should be borne in mind, however, that the meter's accuracy below some −70 dB falls rapidly, as this value approaches the minimum detectable level of the converter chip.

Finally, a switch shifts the decimal point as required, while LED drivers arrange for the relevant unit indication (mV, V, or dB) to light on the front panel.

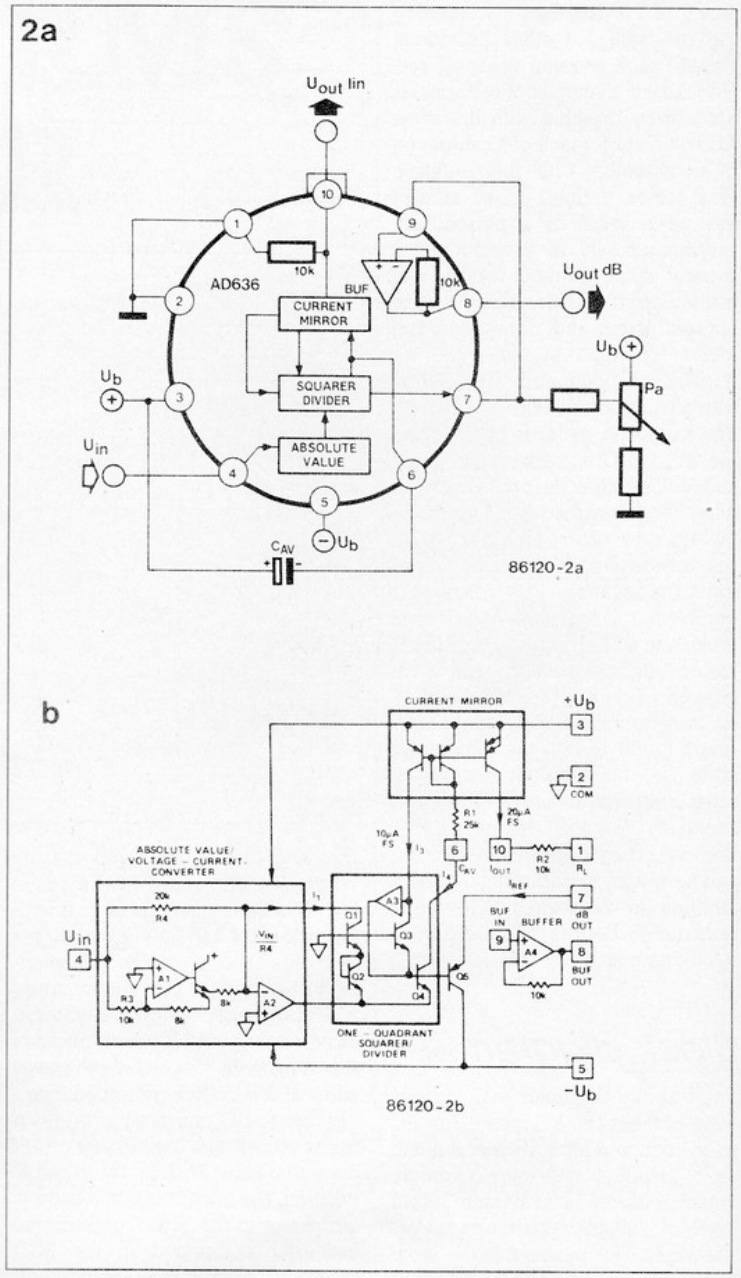


Fig. 2. Internal organization of the Type AD636JH precision RMS-to-DC converter.

RMS-to-DC conversion

In essence, there are two methods for converting an rms level to a proportional DC level, which can subsequently be used to drive a meter indication circuit, whether this be a digital or an analogue type. The first method is based on the use of a thermocouple device, which determines the rms value of measured current or voltage by means of the heating effect in a strip or wire composed of two dissimilar metals joined to form a circuit producing an electromotive force. The second method involves the use of semiconductor devices which incorporate analogue processing circuits for the calculation of a corresponding direct output voltage or output current.

Figures 2a and 2b show what is inside the Type AD636JH rms-to-DC converter chip. It comprises an input rectifier plus voltage-to-current converter, a feedback-current controlled squarer circuit based upon the use of logaritmic and anti-logarithmic amplifiers, which are used to output the logaritmic DC level.

The squared signal is averaged by means of a low-pass R-C network, of which the capacitor, C_{av}, is connected as an external part. The averaged value is converted to a proportional direct current by means of a current-mirror, which passes its output through an on-chip, high-stability, 10K resistor. The proportional direct output voltage is available at chip pin 10.

External preset P_a provides the bias for the internal squarer/divider, and hence can be used to effect calibration of the AD363 at 0 dB input level.

Conversion errors

It stands to reason that any type of practical rms-to-DC converter inevitably produces a small deviation from the ideal conversion characteristics. The main errors and their possible cause will be discussed briefly in the following points.

Static error. Production tolerances and deviations from the target specifications amount to an acceptable level of 1 mV in the case of the

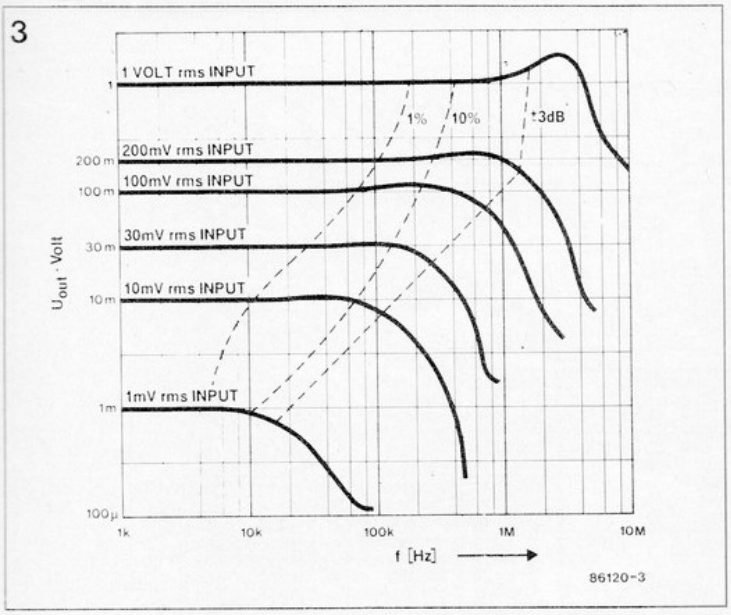


Fig. 3. Converter direct output voltage as a function of input signal frequency, with six rms input voltage levels as parameters.

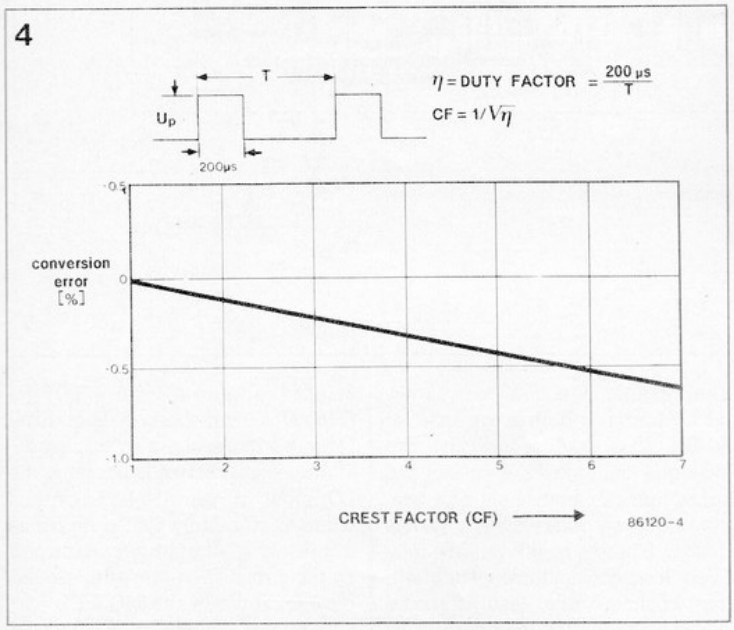
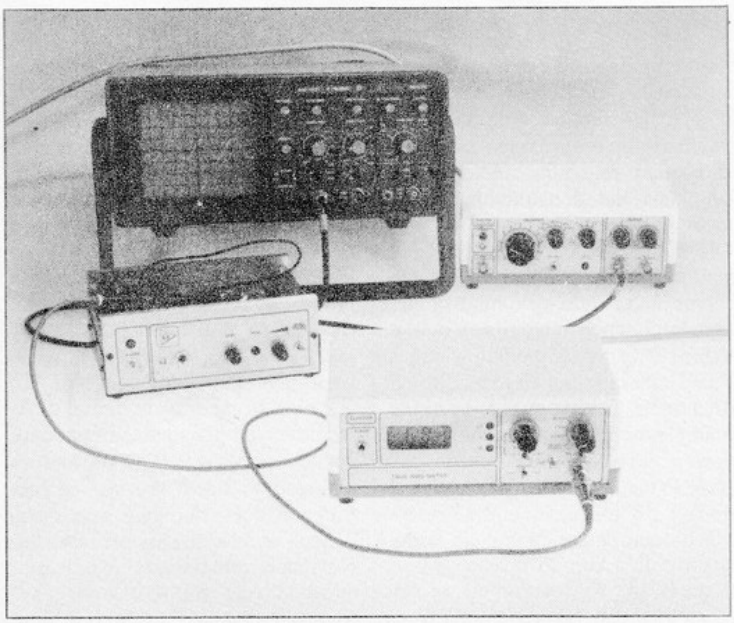


Fig. 4. Correlation between input signal crest factor and conversion error of the Type AD6363JH.



stated converter chip.

Bandwidth. There is, unfortunately, a limitation imposed upon the achievable bandwidth of the converter chip.

Fig. 3 shows the correlation between input signal frequency and the chip output voltage. Note that the converter's usable bandwidth is strongly dependent on the level of the applied input voltage. It is, therefore, advisable to carry out measurements in the lowest possible meter range.

DC error. It is readily understood that C_{av} determines the lowest input signal frequency that produces a faithful direct output voltage; the capacitance of C_{av} therefore requires due consideration in designing with the AD6636JH. In the proposed meter, provision has been made to select one of two capacitors C_{av} to achieve an indication response as required by the specific input frequency.

Crest factor. As already stated in Table 1, the crest factor of a rectangular wave is inversely proportional to its duty factor. Fig. 4 shows the conversion error percentage as a function of the crest factor. The cause for this error lies in the fact that, in the case of very low duty factors (needle pulses), C_{av} has the daunting task of instantaneously "catching" all the energy contained in the pulse, and retain its charge for the averaging process to be completed. Obviously this is very difficult to achieve in practice, whence the relatively small error, which, however, becomes the more manifest when added to the previously mentioned errors, especially when reading rms values of signals with a very high crest factor (i.e. low duty factor).

A special difficulty may arise if a high crest factor signal causes the meter's input section, and hence the converter chip, to be overdriven, since the resulting waveform distortion (clipping) and the generation of harmonics readily leads to erroneous display readings. It is, therefore, suggested to first measure the peak value of such signals using an oscilloscope, to decide on the correct input range of the true-rms meter.

Circuit description

Display unit (see Fig. 5).

The LC display driver/A-to-D converter is a conventional design based on the Type 7106, whose operational details have been covered in *Electronic balance*, elsewhere in this issue. Fig. 6 shows how circuit

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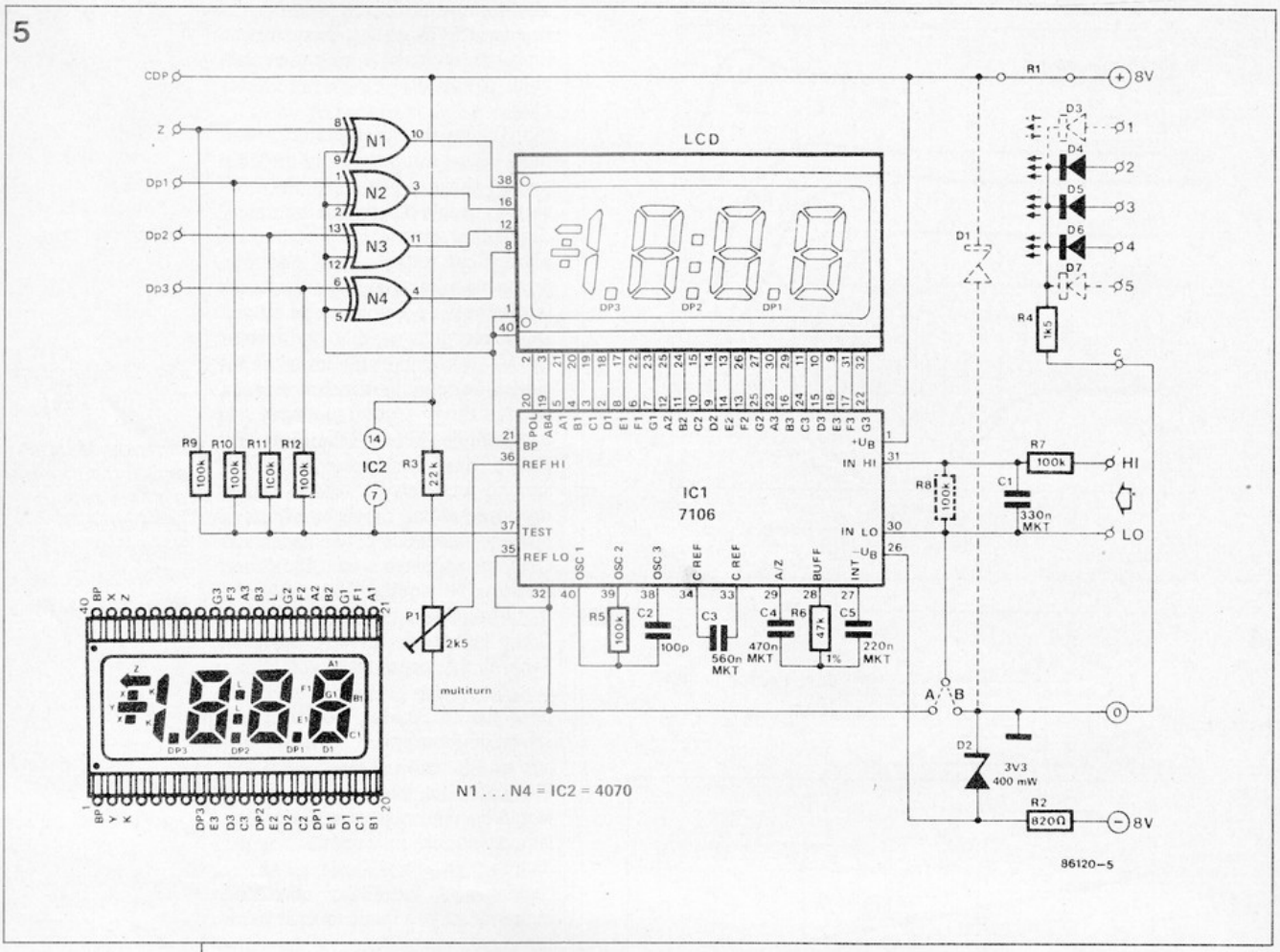

Fig. 5. Circuit diagram of the universal LC display board. Please note that dotted components, including wire A or B, should not be used.

Parts list (display board)

(Note: parts indication is to BS 1852; see Infocard 500)

Resistors:

(tolerance is ±5% unless otherwise stated)

R_2 = 820R
R_3 = 22K
R_4 = 1K5
R_5; R_7; R_9 . . . R_{12} = 100K
R_6 = 47K0F
R_8 = not required
P_1 = 2K5 multiturn preset

Capacitors:

C_1 = 330n
C_2 = 100p ceramic
C_3 = 560n MKT
C_4 = 470n MKT
C_5 = 220n MKT

board 84012-2 is to be completed to make the LC display unit. Point C should be wired to ground, while the dotted components are not required.

Converter board (see Fig. 7).

With reference to circuit diagram Fig. 7, the input voltage enters the meter via a AC/DC selector, S_2/C_A. Next comes a three-resistor, frequency-compensated voltage divider, R_1-R_2-R_3. Two pairs of FETs, T_1-T_2 and T_3-T_4, have been connected to function as ultra-low leakage protective diodes on the 20 mV/200 mV and 2 V input lines.

Low-noise opamp IC_1 is the 10× amplifier for the 20 mV (AC only) range. With the range selector set to 2 V or 20 V, T_5 takes the non-inverting input of the LF356 to ground so as to prevent noise or cross-talk from being amplified. Also Re_1 is de-activated to avoid the meter input impedance from falling to about 10 k (R_4).

Selection of the relevant input voltage range is accomplished with a dual analogue multiplexer. Depending on the binary code applied to its channel selection inputs, pins 9 and 10, each section of the IC passes the analogue voltage at the relevant input Ø3 to the chip output.

The channel selection code is obtained from two-pole range selector S_{6a}-S_{6b}. It is readily seen that the analogue multiplexer is in fact the semiconductor equivalent of a two-pole, four-way rotary switch. In this design, where signal levels are relatively low, conventional switch wiring would readily lead to noise being picked up by long cable runs connected to circuits with a high input or output impedance. It was, therefore, deemed practical to leave all signals "on board" and to accomplish selection with an electronic device ensuring low cross-talk and good reliability. Moreover, it simplifies the construction of the meter to a considerable extent.

The upper section of the analogue multiplexer drives buffer IC_3, which serves to provide matching of the high-impedance multiplexer output to the converter chip input, which is stated to have an impedance of about 6700 ohms. The lower section of the multiplexer provides selection of the correct compensation voltage applied to the display driver's LO input during dB measurements. The requisite compensation voltage is derived from a tapped resistor ladder network at the four inputs of the lower section of IC_2. Switch S_3 selects between a fixed 0 dB level (775 mV_{rms}) and a user-defined level (offset) brought about by turning P_1. During linear rms measurements, the LO input of the display driver is taken to ground by T_6; R_{24} functions to prevent IC_2 from being damaged by the virtual short-circuit to ground brought about by the MOSFET.

Operational amplifiers A_2 and A_3 form the overflow and underflow detector, respectively. Should the voltage at the output of A_1 be higher than 140 mV (8 V/(R_{25}/R_{26}), A_2 toggles and drives the HI line with about 8 V, causing the overflow sign to appear on the LC display. Similarly, A_3 signals underflow if the output of A_1 drops below −300 mV. Diodes D_3-D_4-D_5 constitute an OR gate to disable A_3 from detecting underflow in the −40 dB range.

Opamp A_1 has been included to enable the rms converter output to be temperature compensated. To this end, the feedback circuit of A_1 includes a negative temperature coefficient resistor (NTC), which is arranged to be in thermal contact with the converter chip enclosure. Voltage divider R_{22}-R_{23} provides the previously mentioned 1 mV/dB gradient for the A-D converter contained in the LCD driver, IC_1 in

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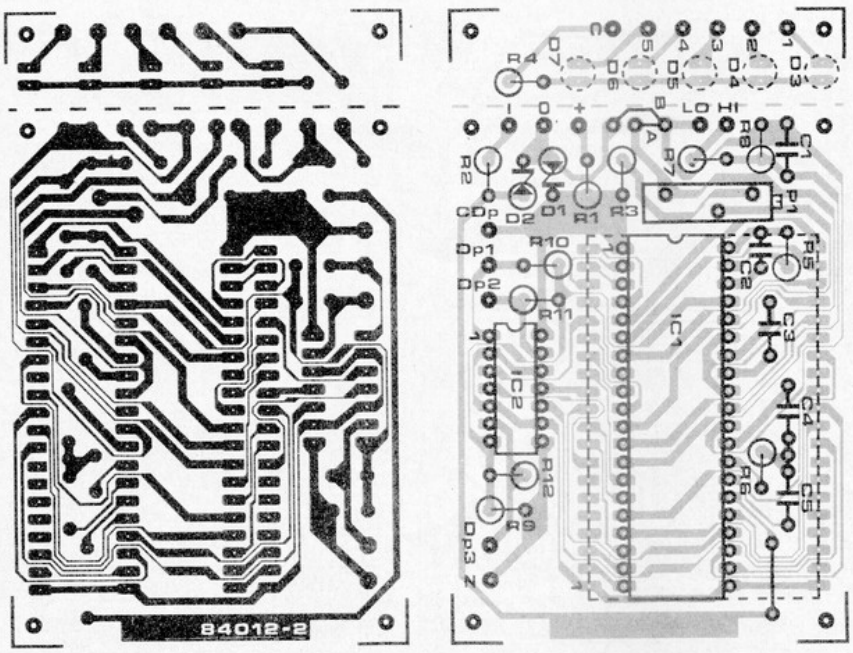

Semiconductors:

D_1 = not required
D_2 = 3V3; 400 mW zenerdiode
$D_4 \ldots D_6$ incl. = red LED (D_3 & D_7 not required)
IC_1 = 7106
IC_2 = 4070
LCD = 3½ digit liquid crystal display; digit heigth 13.3 mm (e.g. Hamlin Type 3901 or 3902 SE 6902).

Miscellaneous:

PCB Type 84012-2 (see Readers Services)

Fig. 6. Track layout and component mounting plan of the display board.

Fig. 7. Circuit diagram of the main converter board.

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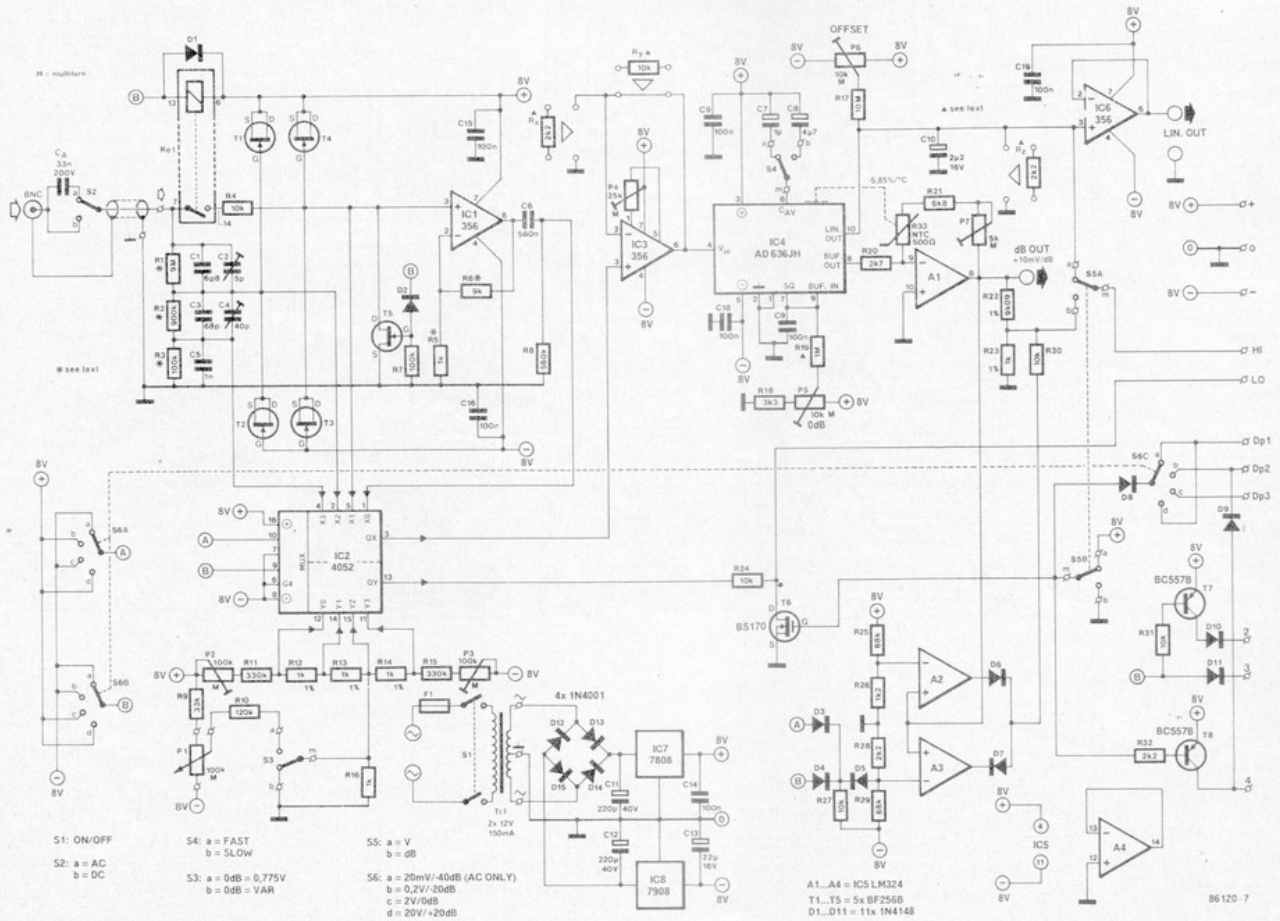
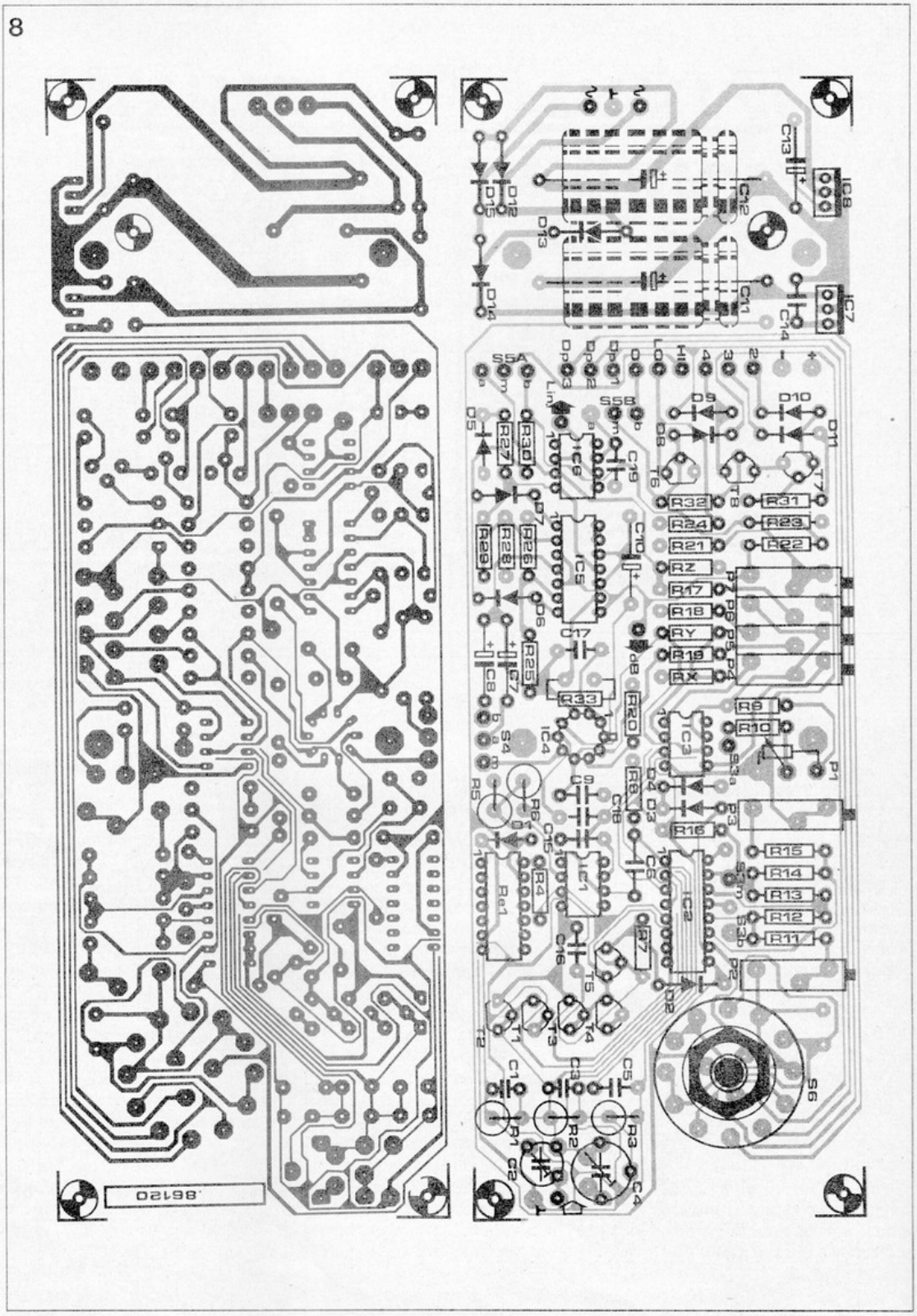


Fig. 8. Track layout and component mounting plan of the main meter PCB. Note that it is double-sided, but not through-plated. C_{11} and C_{12} are to be fitted onto the track side, while a number of components are mounted vertically and with their leads soldered at both PCB sides. The input attenuator screening has been enhanced with the addition of single-ended tracks.



Parts list
(main meter board)

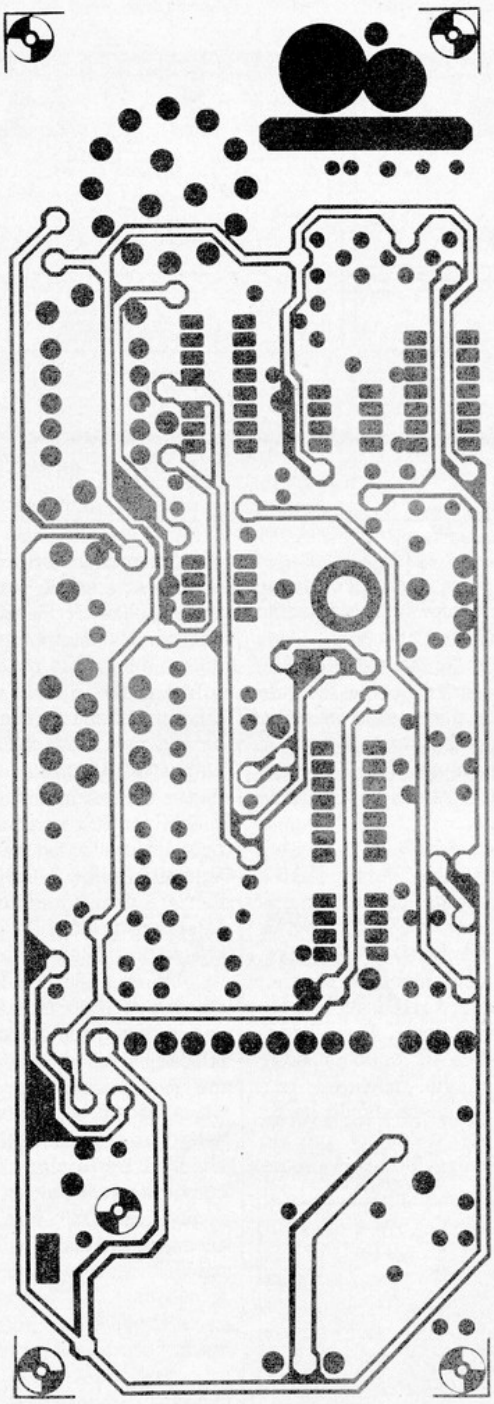
(Note: parts indication is to BS 1852; see Infocard 500)

Resistors:

(tolerance is ±5% unless otherwise stated)

R_1 = 9M; 0.1% *
R_2 = 900K; 0.1% *
R_3 = 100k; 0.1% ▼
R_4;R_{24};R_{27};R_{30};R_{31} = 10K
R_5 = 1K0; 0.1% ▼
R_6 = 9K; 0.1% ▼
R_7 = 100K
R_8 = 560K
R_9 = 33K
R_{10} = 120K
R_{11};R_{15} = 330K
R_{12};R_{13};R_{14};R_{23} = 1K0F
R_{16} = 1K0
R_{17} = 10M
R_{18} = 3K3
R_{19} = 1M0
R_{20} = 2K7
R_{21} = 6K8
R_{22} = 9K09F
R_{25};R_{29} = 68K
R_{26} = 1K2
R_{28};R_{32} = 2K2
R_{33} = NTC 500R; −5.9%/°C; e.g. Mullard Type 2322 610 12501 (±10%) or Type 2322 610 11501 (±20%). (STC; (0279) 26777)
P_1 = 100K linear potentiometer; multiturn.
P_2;P_3 = 100K multiturn preset
P_4 = 25K multiturn preset

Fig. 5.
The meter has two optional outputs, one linear (IC_6), and one logarithmic (full output of A_1). The former may be used to drive an analogue meter in order to observe the trend of the measured voltage; the latter is especially useful for swept-frequency measurements, where an oscilloscope can be used to display curves with amplitude readings given direct in decibels.
Resistors R_X, R_Y and R_Z (denotation ▲) may be fitted to enhance the meter's response to input frequencies in excess of about 100 kHz. Also, R_{19} is changed to 220K. Referring back to Fig. 3, it is seen that the converter chip sensitivity begins to fall appreciably at that frequency, when driven with an input signal of the order of 1-10 mV. Where this is considered problematic, IC_3 may be configured as shown to achieve an amplification factor of about ten (10K/2K2). This modification implies that the converter chip is driven with a higher input signal level so as to improve upon its response to relatively high signal frequencies. However it should be borne in mind that it also implies overdriving the chip, since it receives about 1 V_{rms} rather



than 200 mV_{rms}, which is stated to be the level for optimum accuracy of conversion. The 2K2 resistor, R_2, at the LIN OUT line has been included to keep the display from reading a five times too high rms level.

Finally, the decimal point selection with S_{6c}, and the three-LED range indication, are circuits with a simplicity to obviate the need for any further detailing.

Construction

Constructors wishing to make their own PCB for this project are advised not to use the usual aerosol pc board lacquer, since this material may cause stray resistance in high-impedance circuit sections. The use of plastic or poly-urethane spray may be resorted to, but the best solution in all cases is to order the Type 86120 ready-made PCB from our Readers Services, since this has a protective film to ensure rapid soldering as well as extremely high electrical isolation between tracks and ground plane.

Fig. 8 shows the location of the various parts on the converter board. All parts are fitted in the normal manner; some, however, should have their leads soldered at both PCB sides to effect through-plating. Capacitors C_{11} and C_{12} are fitted at the track side. The NTC, R_{33}, should be fitted on top of the converter chip enclosure, using a small amount of heat conductive paste, and taking due care to prevent short-circuits between NTC leads and the IC can.

All resistors marked with an asterisk are high-stability, 0.1% tolerance types. Replacing these with more readily available 1% types is feasible at the cost of a corresponding loss of accuracy.

PCB holes remaining empty after fitting all parts should be through-plated with pieces of left-over component wire. Do not forget to fit R_y or a wire jumper.

The display board is mounted onto the main converter board by means of four spacers whose length allows the face of the LC display to be level with the enclosure front panel. Use isolating washers or nylon screws and nuts to secure the boards in a sandwich construction. Prior to fitting them onto the board, soldering pins should be cut to a length of 3 mm to preclude short circuits. In the usual manner, the fuseholder, mains socket, and the linear and logarithmic outlets are fitted onto the rear panel. The photograph on page 40 of this article further illustrates the construction of the meter. The studs in the lid of the Verobox enclosure should be removed to enable the unit to be closed properly.

closed properly.

The inside of the enclosure should be lined with aluminium foil to effect screening against stray inductive fields, which would otherwise cause erroneous readings. It is also possible to spray the inside of the box with conductive lacquer. Whichever screening is used, do not forget to connect all surfaces to ground, and beware of short-circuits.

In order to preclude earth loops from being made by the use of the mains earth line, this should not be connected to the circuit ground. Provided due care is taken in the isolation of terminals and wires at mains potential, no problems are expected to arise from the absence of an earth connection.

P_5;P_6 = 10K multiturn preset
P_7 = 5K multiturn preset

Capacitors:
C_A = 33n; 200 V (not mounted on PCB)
C_1 = 6p8 ceramic
C_2 = 5p5 foil trimmer (grey)
C_3 = 68p NP0
C_4 = 40p foil trimmer (violet)
C_5 = 1n
C_9;C_{14} . . . C_{19} incl. = 100n
C_7 = 1μ0; 25 V tantalum
C_8 = 4μ7; 25 V tantalum
C_{10} = 2μ2; 16 V tantalum
C_{11};C_{12} = 220μ; 40 V
C_{13} = 22μ; 16 V

Semiconductors:
D_1 . . . D_{11} incl. = 1N4148
D_{12} . . . D_{15} incl. = 1N4001
T_1 . . . T_5 incl. = BF256B
T_6 = BS170
T_7;T_8 = BC557B
IC_1;IC_3;IC_6 = LF356
IC_2 = 4052B
IC_4 = AD636JH (Electromail; (0536) 204555)
IC_5 = LM324
IC_7 = 7808
IC_8 = 7908

Miscellaneous:
F_1 = 100 mA delayed action
S_1 = DPDT mains switch
S_2 . . . S_4 incl. = miniature SPDT switch
S_5 = miniature DPDT switch
S_6 = 3-pole, 4-way rotary switch
Tr_1 = 2 × 12 V; 150 mA
Re_1 = 15 V DIL reed relay
Enclosure Verobox Type 75-01411D
PCB type EPS 86210 (see Readers Services)
Front panel foil Type 86120-F (see Readers Services)
Output sockets (LOG and LIN) if required

* Available from Trafo-Löwe Elektronik, Postfach 2150, Issum 2, Sevelen, West Germany. Telephone: +49 2835 5012/5013; Telex 08 12 261.

▼ Welwyn RC55 series 0.1%, 15 ppm; or Rhopoint Econistor/Miniohm series (10K; 9K; 100K; 1K0) 3 ppm. Both series available from STC Electronic Services, (0279) 26777 (note: Welwyn series only 10+ !).

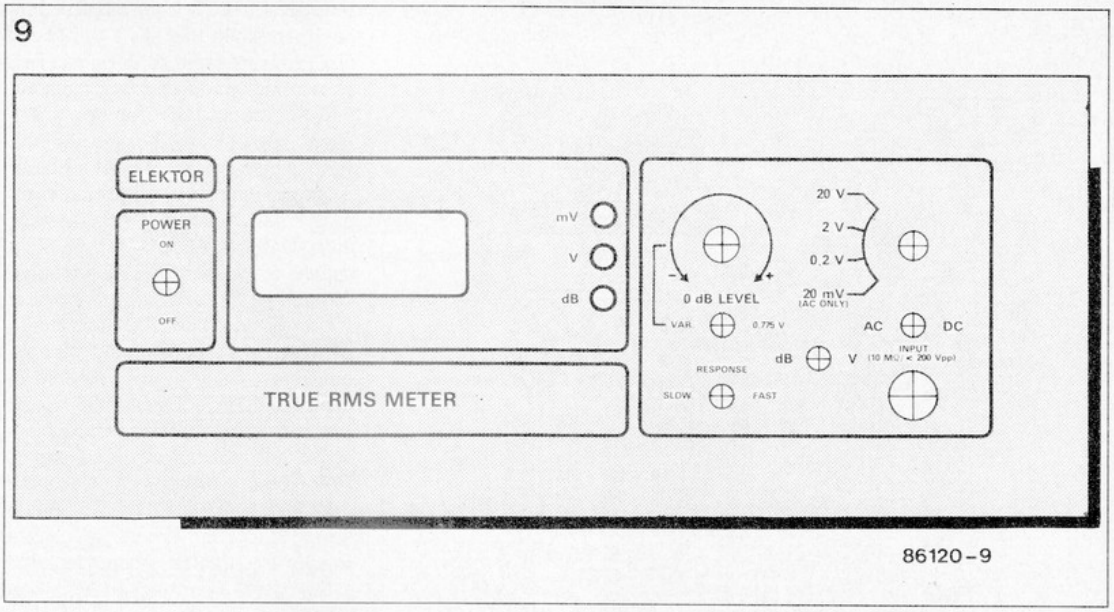


Fig. 9. Front panel foil for the true-RMS-meter.

Setting up

Before embarking on the setting up of the instrument, it is suggested to leave it switched on for about 20 minutes so as to ensure sufficient thermal stability.

To begin with, the linear (V) functions are adjusted as follows. Short-circuit the meter input, set S_2 to DC, and set S_6 to the 200 mV range. Adjust P_4 for 0 mV with respect to ground, measuring at the output of IC_3. Next, adjust P_6 for a display reading of 00.0. Apply a direct voltage of 150 mV to the meter input and turn P· on the display board for a display indication of 150.0.

Proceed with the dB functions of the meter by first setting S_5 to the dB position; S_2 is left in the DC position, and S_6 is set to the 2 V (0 dB) range. Select the fixed 0 dB level with S_3. Apply a direct voltage of 77.5. mV to pin 3 of IC_3 and set S_5 for a display reading of 0 dB. Next, select the +20 dB range (20 V) and set P_3 for a display reading of 20.0. Verify that the meter reads −20.0 when set to the −20 dB (0.2 V) range; if necessary, correct the indicated value by means of adjusting P_2. Finally, apply a direct voltage of 77.5 mV to the meter input, select the 0 dB range (2 V), and adjust P_7 for an LCD indication of −20 dB.

The input attenuator is preferably aligned using an oscilloscope and a generator capable of supplying a rectangular wave of 1 kHz at 1 V_{tt} and 10 V_{tt}. Apply the 1 V_{tt} square wave to the meter input and connect the scope to pin 6 of IC_3. Set S_6 to the 2 V range, and S_2 to DC. Carefully adjust trimmer C_2 for optimum edge steepness of the displayed rectangular signal; there should be no undershoot or overshoot on the leading or trailing edges. Increase the generator output voltage to 10 V_{tt} and select the 20 V meter range; peak C_4 like C_2. Redo the trimmer adjustments until both meter ranges offer a satisfactory pulse response.

In the absence of an oscilloscope, the input attenuator may be aligned using a sine wave oscillator whose output voltage is known to be accurate. Set the generator to produce a 10 kHz, 1 V_{rms} sine wave, and connect its output to the meter input. Select meter range 2 V, DC, and adjust C_2 for a display reading of 1.000 V. Select meter range 20 V, DC, and increase the generator output voltage to 10 V_{rms}. Adjust C_4 for a display reading of 10.00 V. Switch between the two ranges and each time carefully adjust the relevant trimmer until the meter reads the correct rms value of the applied voltage. Finally, the attenuator alignment may checked by varying the generator output frequency to see whether the display indication remains in accordance with the set sine wave amplitude.

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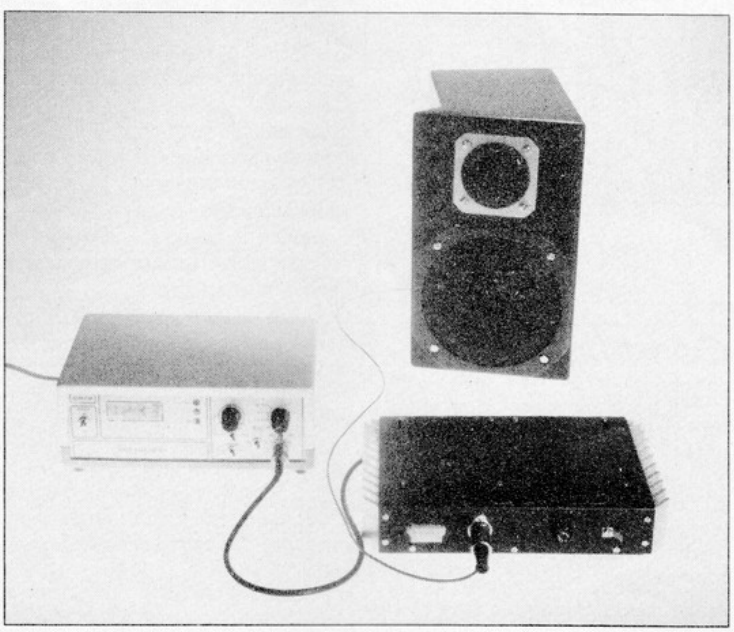


ANALOGUE WATTMETER

by C Giorden

This simple meter is primarily intended to establish how much power a mains load consumes and thus how much it adds to the electricity bill.

We all know that the power consumption of electric appliances is generally stated in watts, and that the cost of using the appliance is roughly proportional to its power drain from the mains. It would appear interesting to investigate the phenomenon "power" a little further, particularly since the simple notion that power is the product of current and voltage ($P = I \cdot U$, whence $P = U^2/R$ and $P = I^2R$) is not directly applicable to alternating voltage.

A brief summary of terms may help to shed some light on the operational principle of the proposed wattmeter. Simply measuring the rms values of alternating current and voltage developed across a load R yields the *apparent power*: $P = I_{rms} \cdot U_{rms}$ [W]. However this method can only be used where the load R is purely resistive; should it consist of a reactive part (capacitance, inductance, or both) and a resistive part, as is the case with most mains loads, the calculation becomes more complex, since only that component of the current which is in phase with the voltage adds to the *active power* consumed by the load. The greater the phase shift, expressed as an angle, φ, between the voltage across, and the current through, the reactive load, the lower the active power, whence $P = U_{rms} \cdot I_{rms} \cdot \cos \varphi$ [W]. For purely resistive loads, apparent power equals active power, since there is no phase shift, and hence $\cos \varphi = \cos 0° = 1$.

Our interest, naturally, lies in establishing the amount of active power consumed by an electric appliance, since that quantity, integrated over the on-time, is faithfully recorded by the ever ticking kilowatt-hour (kWH) meter fitted by the local electricity board.

A more extensive discussion of theoretical aspects involved in power consumption can be found in *What is power?*, *Elektor India*, June 1983, p. 6-29 ff.

Wattmeters

All roads lead to Greenock, Renfrewshire, the birthplace of James Watt. However the most commonly used type of wattmeter is the electrodynamometer, in which a fixed coil carries the measured current, while a coupled moving meter coil accepts the alternating supply voltage via a series resistor. The meter deflection produced in this instrument is proportional to the wattage of the load, whether this be a pure resistance or a combination of resistance and reactance. In essence, the electrodynamic wattmeter provides the average value of the instantaneous product of rms current and rms voltage. It is recalled that the term "instantaneous" relates to the previously mentioned phase shift, φ, between U_{rms} and I_{rms}.

In an electronic wattmeter, one would need to measure the rms (effective) value of both voltage and current, as well as the phase difference, if present. Next, these three data would have to be multiplied to arrive at a possibly close approximation of the active power figure. Two difficulties may crop up in this concept: one concerns the rms calculation for the specific waveform (see *True RMS meter*, elsewhere in this issue); the other is the practical realization of a phase difference measuring circuit which outputs the cosine factor, φ.

The proposed analogue wattmeter goes round these problems since it may be viewed upon as an electronic version of the electrodynamic wattmeter, using the same underlying principle of measuring the instantaneous rms values of current and voltage, and multiplying these prior to time-averaging the product. The final version of the analogue wattmeter is not intended to pass for a superbly accurate instrument; rather, we set out to keep it cheap, passive (i.e. it requires no external supply), and simple to construct.

A multiplier circuit

Fig. 1 shows the basic circuit of a voltage-current multiplier. Initially, current I_2 is assumed nought. Provided the transistors have identical characteristics, their base-emitter voltages, and hence their collector currents, are equal. Preset P is used to level out small differences between $I_{c(T1)}$ and $I_{c(T2)}$ due the production tolerance of the transistors. The circuit as shown is often referred to as a *current mirror*. With $I_2 = 0$ and $I_{c(T1)} = I_{c(T2)}$, there is no current flow, I_3, through the meter coil. However any current I_2 causes a voltage drop, ΔU, across R. The resulting potential difference between the emitters of T_1 and T_2 causes a proportional compensation current, I_3, to flow between the two collectors. Provided the variation in V_{be} of T_2 remains relatively small, there exists, in theory, a linear relation between ΔU and I_3.

The current factor, I in $P = U \cdot I$, is obtained from $I_{c(T1)}$ and $I_{c(T2)}$. The linear relation between I_1 and I_3 is most readily explained in the knowledge that the transistors' mutual conduc-

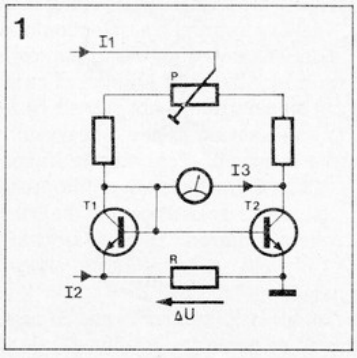
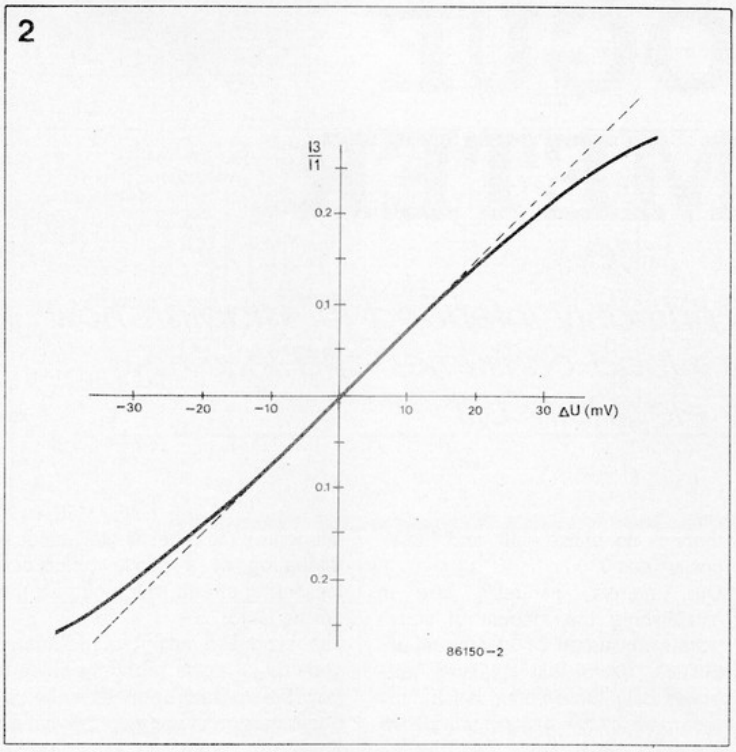


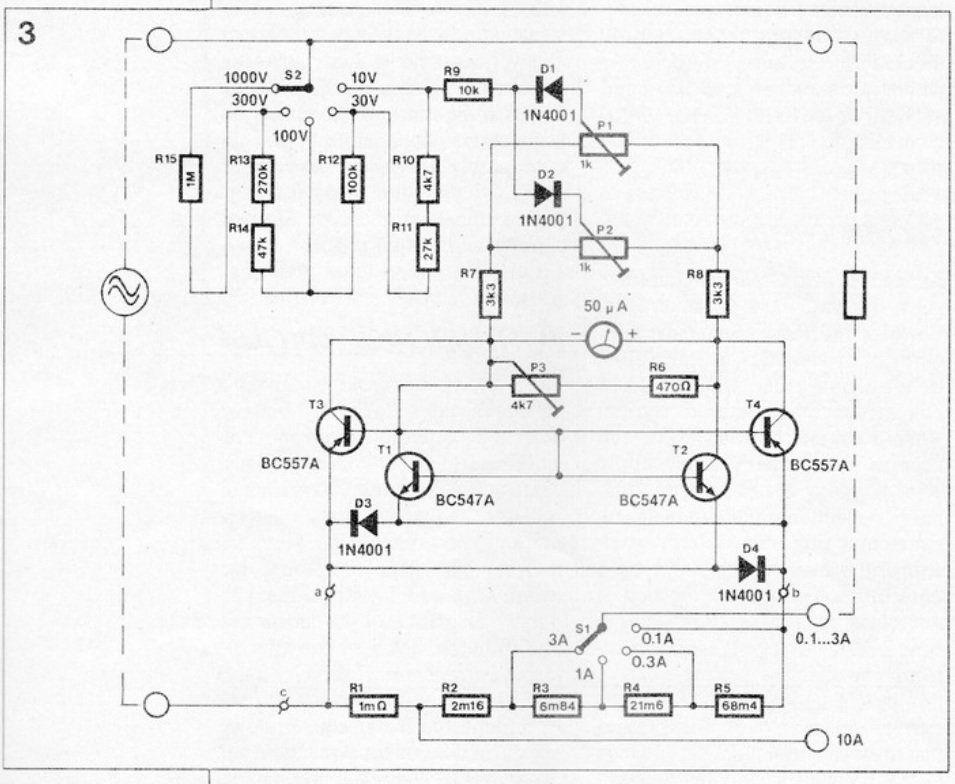
Fig. 1. Basic configuration of a voltage-current multiplier.

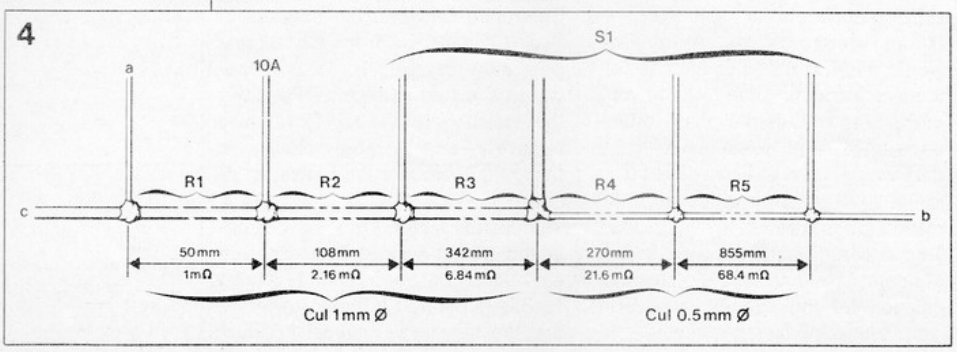
Fig. 2. The curve shows that the multiplier circuit of Fig. 1 operates linearly within 4%, provided ΔU is less than ± 20 mV, and I_3 remains a fraction of I_1.

Fig. 3. The wattmeter essentially comprises two multiplier circuits, i.e. one for each half period of an alternating voltage.

Fig. 4. Shunt resistor network $R_1 \ldots R_5$ incl. is made from enamelled copper wire, which affords good definition of the resistance ratios.







tance ($\Delta I_c/\Delta U_{be}$) is directly proportional to the collector current. With ΔU constant, therefore, doubling I_1 causes a doubling of the meter current, I_3.

The limitations imposed upon the use of the current mirror mainly concern the range of ΔU and the ratio of the quiescent collector current to the meter current. At $\Delta U = 20$ mV, for instance, the non-linearity error amounts to about 4% (see Fig. 2). Non-linear behaviour of the current mirror also originates from dissimilar transistor characteristics, which in practice means that the devices will have to be selected for near identical operation in the circuit. It should be realized that the base-emitter temperature coefficient of about -2 mV/°C at $\Delta I_c = 0$, is not small relative to ΔU, and that, therefore, the transistors should be fitted in close thermal contact.

Circuit description

The foregoing considerations have led to a practical circuit, shown in Fig. 3. In order to enable the taking of AC power measurements, and also to go round a meter polarization circuit, D_1 and D_2 automatically select the relevant multiplier, T_1-T_2 or T_3-T_4. Each of these circuits has its own balance preset, P_1 or P_2. D_3 and D_4 serve to preclude breakdown of base-emitter junctions due to negative voltage surges. The collector-emitter voltage of all four transistors is kept low at about 0.7 V, since the bases of T_1 and T_3 have been connected direct to the collectors. This implies that the voltage drop across the meter coil may not be so high as to cause T_1 or T_3 to be driven into saturation, which would lead to erroneous meter readings. At a coil resistance of some 1 to 1.5 kilo-ohms, the meter drops less than 100 mV, ensuring ample drive reserve for the circuit in the case of measuring high current peaks (AC signals with a relatively high crest factor).

The meter's range selectors, S_1 (A) and S_2 (V) ensure operation of the multipliers within their linear ΔU and I_1 span; the former is kept lower than 20 mV by appropriate selection of a shunt resistor from series network R_1-R_5 incl. In the 10 A range, the output load current is not carried by S_1 to avoid having to use a heavy-duty rotary switch. The voltage range switch, S_2, selects an appropriate total series resistance for the total collector current to be reasonably steady over a large input voltage range.

The resistors around S_1 and S_2 have been selected for a range factor of

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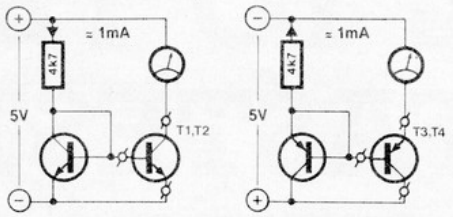

Fig. 5. Basic circuit to select transistors for matching characteristics.

√10 (≈3.16), in order that the wattage can be read from a double meter scale (see Fig. 7). The scale factor also affords readings to be taken without the need for calculations; to this end, the wattmeter's front panel has a U·I product matrix (see Fig. 6).

Construction

The circuit is constructed on a piece of Veroboard. Fig. 4 shows a suggested construction of shunt resistor network R_1 . . . R_5 incl.; it simply comprises two series connected lengths of ϕ 1 mm (SWG20) and ϕ 0.5 mm (SWG24) enamelled copper wire, tapped at suitable locations to effect the correct resistance ratio. The resistivity of ϕ 1 mm enamelled copper wire is about 20 mΩ/m, that of ϕ 0.5 mm about 80 mΩ/m. The tapped shunt wire may be wound onto a suitable former without a core. At about half way of the winding, reverse the winding direction to lower the overall inductance of the shunt. As shown in the circuit diagram, S_1 has been connected such that relatively high currents (3 A, 10 A) are arranged to bypass the high resistance section of the shunt, R_3-R_5. This effectively precludes the shunt wire heating up and thus increasing its resistivity. The contact rating of S_1 should be duly observed; if necessary, use a two-pole rotary switch and connect the poles and contacts in parallel. In general, the shunt section should be made in observance of the relatively high currents and voltages involved.

As already stated, the transistors in this circuit should be closely matched types. Fig. 5 shows test circuits for T_1-T_2 (npn) and T_3-T_4 (pnp), in which the transistors can be examined for equal thermal characteristics. While measuring, make sure that the transistors are in thermal contact, but do not press them together with your fingers.

It goes without saying that T_1-T_4 in the wattmeter should also be fitted in close thermal contact; it is suggested to apply heat conductive compound onto the enclosures before these are clamped together with a strip of brass or copper sheet.

The meter circuit is best fitted in an ABS enclosure as shown in Fig. 6. In view of the possible use of the meter with mains-operated equipment, it is very important to use isolated knobs and good quality switches. Also watch out for the meter adjustment screw, if this is a metal type.

Setting up and use in practice

Apply a direct voltage of about 30 V to the meter input (— to the pole of S_2) and zero the indication. Should this be impossible, T_1 and T_2 are not sufficiently matched, and either one of these transistors needs to be replaced with a better type. This problem may occur in spite of apparently good results in the test circuit of Fig. 5, since finding the collector currents to be equal for one specific bias level need not imply equal characteristics over the whole range of I_C.

Reverse the input voltage polarity and check the performance of T_3-T_4 as with T_1-T_2.

Apply an accurately known alternating voltage to the meter input, and connect a suitably dimensioned load to the meter output. Calculate the wattage, and calibrate the meter with P_3.

Using the meter for DC power measurements requires checking transistor pairs T_1-T_2 and T_3-T_4 for equal characteristics. For a given input voltage and load, reversing the input polarity should not cause different meter readings. However if correct operation is unattainable, fit either another pair T_1-T_2 or T_3-T_4.

While deciding on the appropriate voltage and current setting of the meter, it should be borne in mind that the shunt drop must not exceed 20 mV. This means that, for instance, in the 10 A range, the peak value of the measured current must not exceed 20 A (20 A × 1 mΩ = 20 mV). For direct currents, therefore, the next higher meter range must be selected if the measured current is double the range value. For alternating currents, a factor 1.5 should be observed. The wattmeter's voltage setting is also rather uncritical; the 30 V range, for instance, permits readings to be taken in the 0-100 V_{rms} range.

Finally, since the proposed wattmeter comprises only very few parts, it is well suited for permanent incorporation in appliances such as loudspeakers and light dimmers. The range selectors can, then, be dispensed with, and fixed resistors may be fitted in accordance with the typical power drain of the appliance.

TW

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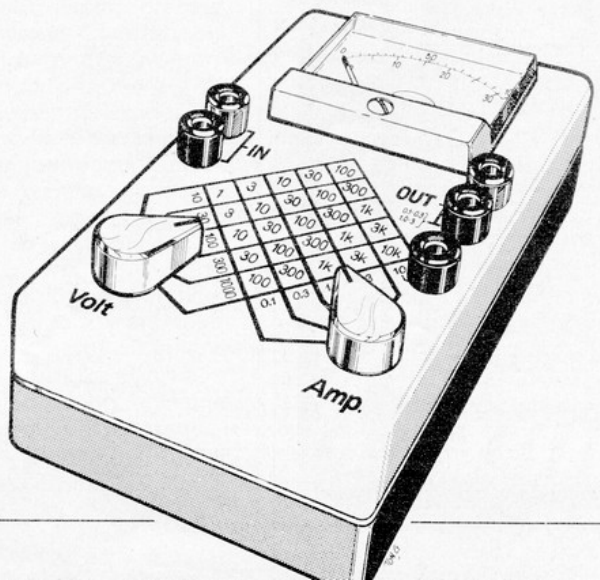

Fig. 6. Artist's impression of the finished analogue wattmeter.

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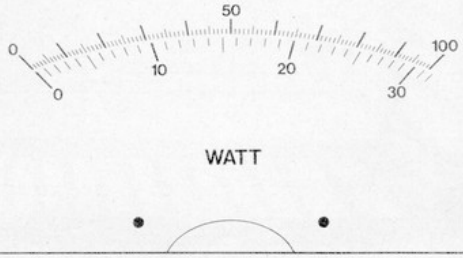


Fig. 7. Suggested scale divisions to match the meter's range factor of √10 (≈3.16).

NEW RIVAL TO THE WORLD'S FASTEST COMPUTERS

by Sandra Smith

Britain staked a large claim in the world of information technology earlier this year with the launch of a computer to rival the fastest in the world. It is Floating Point Systems's [1] new family, the T-series supercomputer.

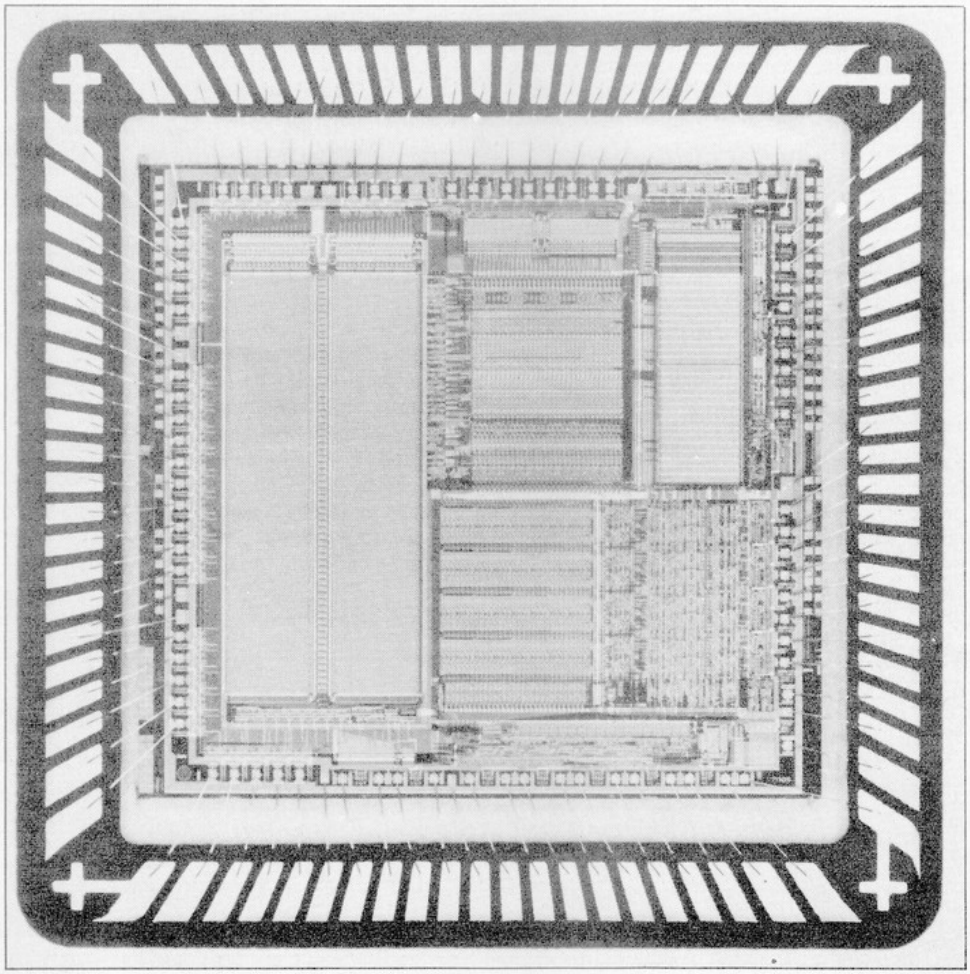
The key to the T-series's unquestionable superiority is the transputer, popurlarly called the computer on a chip. This invention is the brain-child of Ian Barron, one of the founders of Inmos[2], the semiconductor company set up by the Government in 1978. It was established to secure a place for Britain in the mass market for microprocessors and memory chips.

In 1984 Inmos was bought by Thorn-EMI [3]. Its transputer could now become the world's standard chip for supercomputers and Inmos's market value, it is predicted, could rise to at least £200 million by next year on the basis of its success.

The transputer brings nearer the dream of a computer not bigger than a suitcase yet powerful enough to model a nuclear explosion, or to plot a space vehicle's path to distant planets. It can be used to process the massive amounts of information involved in generating and manipulating images in the field known as computer graphics. Even powerful present-day computers can take hours to process graphic images in, for example, television broadcasts, computer aided design, or film animation.

Complex graphics

The transputer can produce complex graphics as quickly as the operator



The Inmos IMS T414 transputer—a computer on a chip.

can think. In use in supercomputers—machines with greater speed, accuracy and memory size and speed than lesser computers—the transputer allows the simulation of experiments formerly possible only in laboratories. This of course makes these processes cheaper than the real thing. The boon to engineers is that they can play "what if?" experiments with proposed designs.

Weather prediction is a classic example of supercomputer application. The idea is to perform the simulation much faster than nature so that one knows the weather before it happens. Even one day forecasts, if sufficiently accurate, can save lives by allowing people to prepare for hurricanes, flooding, and so on.

Another application that "speeds up time" is the simulation of ways to extract oil from proven reserves.

Supercomputers enable constructors to ask questions about complicated structures that require a great deal of planning. The answers enable them to know about stress factors; the use of more suitable materials; aerodynamic design in vehicles; heating and vibration problems and a whole range of essential facts.

Design award

This year, two researchers at University College Hospital, London, won an award for designing a low cost system for modelling the effects of facial plastic surgery using four transputers to build a graphics processor rigged up to the host computer already at the hospital.

It would not be stretching the point to say that the size of projects a country can undertake can be dictated by the maximum speed of its supercomputers. This is why the low-cost transputer is a revolutionary development. The chips cost about £350, and should fall to £50 as production and demand increase.

The T-series's smallest machine, the FPS-264, costs under £350 000—less than a tenth of its nearest rival, the heavyweight Cray [4]. There are only 12 Cray computers in Britain, costing between £10 million and £20 million each. Up to 1000 linked transputers can perform as well as the most powerful Cray.

Comparisons between transputer-based machines and Crays are not reliable, however, as it is not comparing like with like. A yardstick for such comparisons can be the number of floating point operations or arithmetical calculations a supercomputer completes in a second. They are counted in thousands of millions (gigaflops). The Cray 2 offers one gigaflop and costs between £5.2 million and £14.5 million. The T-series's largest machine, the T/40000, has a top speed of 262 gigaflops, about 200 times more than any comparable machine, and it costs only in the region of £1 million.

Concurrent processing

The key to the transputer and the T-series is parallel or concurrent processing, an array of transputers dealing with information in parallel rather than a piece at a time.

Conventional computers process data serially—one task at a time. Parallel machines process several different parts of an operation simultaneously, often millions at a time, by linking a number of processing elements together. This speeds up the process but creates complex design and programming difficulties. The patterns for the inter-linking of the processors determine the complexity of the arithmetic and data manipulations that can be carried out. To simplify the programming of parallel processors a new programming language, Occam, was developed in 1983. According to world transputer expert and co-author of *The Inmos Saga*, Mick McLean: "Like the transputer, the essence of Occam is its simplicity".

Until now the power and versatility of parallel computers has been restricted by the bulkiness of their processors. These are made up of a number of chips, making their ability a trade-off between power and choice of inter-linking pattern. The transputer eliminates these restrictions because it is a complete processing element on a single chip.

Integration techniques

Made by very large-scale integration (VLSI) techniques, it contains the equivalent of 200 000 transistors on a chip which is less than 9 mm² and has a central processor which handles 32 bits of data at a time. It also has a built-in memory and high-speed communication links for exchanging data with other transputers.

Floating Point Systems has chosen a "hypercube" arrangement for its processor inter-linking patterns in which eight transputers link to form the corners of a cube. The hypercube was found to be the most efficient for complex arithmetical calculations.

Each hypercube is called a node, two of which, housed in a filling cabinet-sized container, give as much processing power as a large mainframe computer.

To increase processing power, more hypercubes are simply added, up to a maximum of 2048.

A machine called Computing Surface, designed by Meiko[5], of Bristol will not have such size limitations. Its design will allow it, in theory, to link a limitless number of transputers together. The company's founders are ex-Inmos managers who were involved in the development of the transputer chip.

Recognition systems

Miles Chesney, one of the former managers, says Computing Surface will, for example, be used in "pattern recognition systems for robots and modelling the neural networks of the brain".

The computer has 150 transputers and can handle 1200 million program instructions a second—the sort of performance that only the biggest supercomputers can achieve. Yet it costs only about £250 000, fits into a box the size of a microwave oven, and requires little power.

1. Floating Point Systems (UK) Ltd • Apex House • London Road • Bracknell • Berkshire RG12 2TE.

2. Inmos Ltd • Whitefriars • Lewins Mead • Bristol BS1 2NP.

3. Thorn-EMI PLC • Upper St Martins Lane • London WC2 H9ED.

4. Cray Research (UK) Ltd • Cray House • London Road • Bracknell • Berkshire RG12 2SY.

5. Meiko Ltd • Southgate • Whitefriars • Lewins Mead • Bristol BS1 2NP.

TOP-OF-THE-RANGE PRE-AMPLIFIER

PART 2

This month's instalment of the article deals with the description of the circuits of the phono and line amplifiers. Constructional details will follow in next month's issue.

Initial considerations

In essence, there are only two types of good-quality pick-up cartridge in use nowadays: the moving coil—MC—and the magneto-dynamic—MD. The main difference between these is the level of output voltage they provide. The signal provided by a modern MC cartridge is 100-400 μV, whereas that of a MD type is 2-5 mV.
In the design of an IEC (RIAA) equalization circuit that is suitable for use with both types of cartridge, there are two choices: one in which one or more stages of amplification can be switched on or off depending on whether an MC or an MD cartridge is used; or one with variable (or switch-selected) gain. In the present preamplifier, the second choice has been adopted.
This choice makes heavy demands on the input stage, because it is not easy to achieve a good signal-to-noise ratio with MC cartridges owing to the combination of low output voltage and small hum resistance of these elements. At very low hum impedances (a few ohms), the noise of the input stage tends to drown the signal. Fortunately, careful design and the use of specially selected components can reduce the noise factor of the input stage to a very low value.
Another requirement of a universal input stage is that its input capacitance and resistance can be varied: some MC cartridges should be terminated into 47R, others into 100R, while MD cartridges need a much higher resistance, around 47k. The input capacitance is particularly important when MD cartridges are used, because it affects the frequency response of these elements between 10 kHz and 20 kHz.
The IEC de-emphasis characteristic is obtained by a well-tried combination of a passive low-pass filter and an active low-frequency correction section as shown schematically in Fig. 8. Input stage A_1 raises the level of the signal from the cartridge: its amplification is matched to the type of cartridge and the input voltage by switched resistors.
The output of A_1 is passed through a passive low-pass section which has a cut-off frequency of 2120 Hz. The signal is then applied to amplifier A_2, the negative feedback loop of which contains a low-frequency equalization section that has cut-off frequencies at 500 Hz; 50 Hz; and 5 Hz.

Circuit of the phono amplifier

The input capacitance and resistance are selected by DIL switch S_2—see Fig. 9. Capacitors C_5, C_6, and C_7 are necessary to prevent any DC from reaching the sensitive MC cartridge. All capacitances in the signal path are formed by parallel combinations of a polypropylene and a polyester capacitor: this will be further gone into in Part 3.
The input amplifier consists of three dual transistors Type MAT-02. These are low-noise, carefully matched devices with a low offset voltage drift—see Table 1. Current source T_4 provides the DC bias for the transistors. The voltage drop across a LED is used as the reference voltage. The three transistors are connected in parallel because the base resistance of the input transistor generates most of the thermal noise when the signal source has a very low output impedance: the thermal noise here is, therefore, reduced by nearly 67%.
Another type of noise, the so-called Schottky noise, is determined largely by the collector current of the input transistor. Generally, the Schottky noise diminishes when the collector current increases—up to a limit. The collector current here is set at 1 mA per transistor; ideally, it should have been 3 mA (according to the manufacturers), but this value would give difficulties with the active offset control, which will be discussed later. A value of 1 mA is, however, a good compromise, resulting in excellent signal-to-noise ratios.
The value of input capacitors C_5, C_6, and C_7 also has an effect on the amount of noise. The total value should be of the order of 100 μF to 200 μF for a negligible contribution to the noise at the lowest frequencies. Since electrolytic capacitors can not be used in the signal path (because of their poor performance), a compromise was found empirically between the dimensions of the capacitors and their noise contribution.
All the measures to reduce the noise to an absolute minimum are for the benefit of MC cartridge users: if only MD pick-ups are used, one MAT-02 is sufficient. The collector current of that one transistor can then be reduced to, say, 560 μA by increasing the value of R_{29}.
The input stage forms one half of a differential amplifier: the other half is IC_1. This opamp raises the level of the difference signal at the collectors of the dual transistors.
Opamp IC_1 is a high-quality device

that is not cheap but which gives an excellent performance: its electrical characteristics are given in Table 2. The negative feedback loop of this stage—R_{14}-R_{13}-R_{12}-R_{17}—contains two switches, S_1a, S_1b, with which the input sensitivity can be set to 0.1 mV; 0.2 mV; 2 mV; and 4 mV. This arrangement enables optimal matching of the dynamic range and the signal-to-noise ratio to the output signal of the cartridge. Note the low values of R_{12} and R_{17} which enable the noise at the inverting input of IC_1 to be kept to a minimum.

The difference in value between R_5 and R_{12}-R_{17} results in a relatively large, unwanted offset voltage. This is particularly troublesome when an MC cartridge is used, because the gain of the input stage is then between 40 dB and 46 dB. This problem is resolved by integrator IC_3, which provides active offset correction. The output of IC_1 is first taken through a low-pass filter, R_{25}-C_{16}, with a cut-off frequency of 0.3 Hz, and then integrated by IC_3 . The DC level at the bases of the dual transistors is set by IC_3, via R_{15} or R_{16} depending on the position of S_1b, to a value which results in zero output of IC_1.

Since the supply voltage to a Type LF411 should not exceed 30 V, a 6k8 resistor—R_{35}—has been inserted in the positive supply line to reduce the supply to IC_3 to about +10 V. This creates no problems, since the output of the opamp is always negative when the offset is being adjusted: the output current then flows via the negative supply line.

The current required from IC_3 is fairly large, mainly because of the low value of R_{17} (even a small potential difference across this resistor requires a fairly large current). The output current must be 6—8 mA to keep the output voltage at zero. This explains why the collector current of the dual transistors is arranged at 1 mA: higher values would necessitate an even larger current through R_{12}-R_{17}. Increasing the values of the feedback resistors would result in increased noise in the input stage.

The passive part of the de-emphasis circuit is formed by R_{18}-C_9-C_{10}: the capacitors are 1% polystyrene types. The output of IC_1 is fed to the non-inverting input of a second Type OP-27 amplifier. The negative feedback loop of this stage contains the low-frequency correction section of the de-emphasis circuit. All resistances in the loop are formed by two

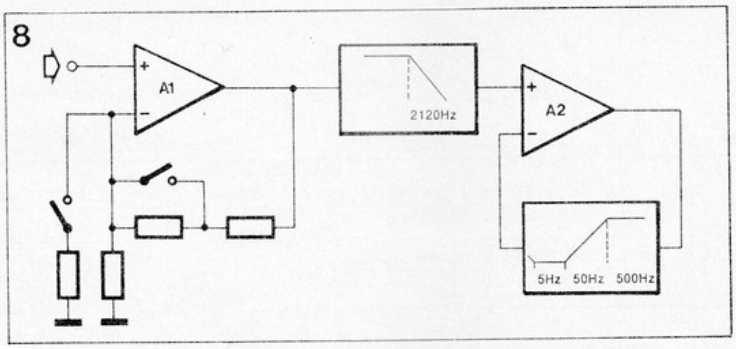


Fig. 8. The twofold nature of the phono amplifier becomes clear from this block diagram.

Fig. 9. The circuit diagram of the phono amplifier. Special low-noise transistors are used in the input stage. Great care has been taken in the design of the IEC (RIAA) equalization circuit, which is split in a passive and an active section.

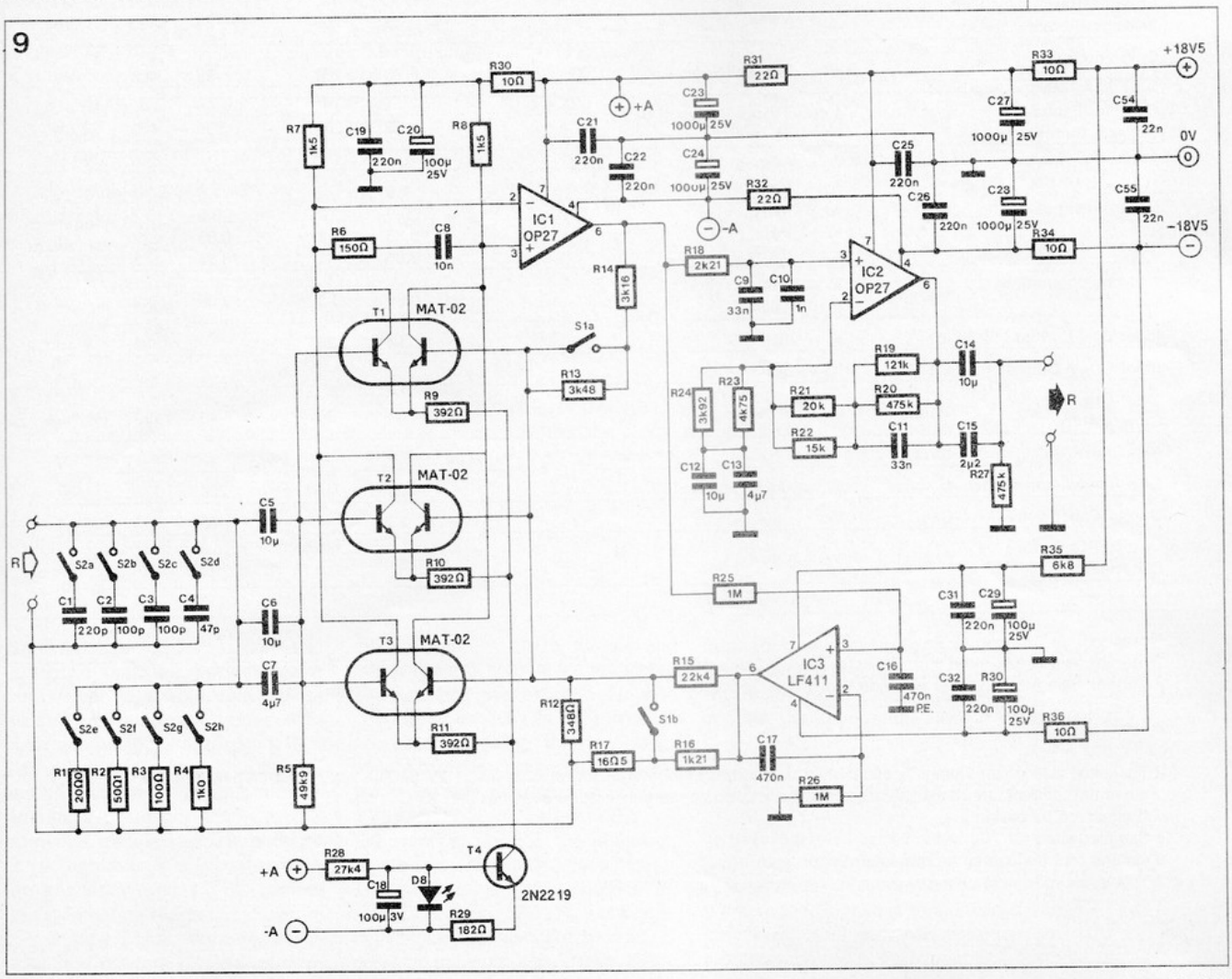


Table 1. Main electrical characteristics of the MAT-02 dual transistor.

Table 1.

V_{CB} = 15V, I_C = 10μA, T_A = 25°C, unless otherwise noted.

PARAMETER	SYMBOL	CONDITIONS	MAT-02A/E MIN	MAT-02A/E TYP	MAT-02A/E MAX	MAT-02B/F MIN	MAT-02B/F TYP	MAT-02B/F MAX	UNITS
Current Gain	h_{FE}	I_C = 1 mA (Note 1)	500	605	—	400	605	—	
		I_C = 100μA	500	590	—	400	590	—	
		I_C = 10μA	400	550	—	300	500	—	
		I_C = 1μA	300	485	—	200	485	—	
Current Gain Match	Δh_{FE}	10μA ≤ I_C ≤ 1mA, (Note 2)	—	0.5	2	—	0.5	4	%
Offset Voltage	V_{OS}	V_{CB} = 0 1μA ≤ I_C ≤ 1mA	—	10	50	—	80	150	μV
Offset Voltage Change vs V_{CB}	$\Delta V_{OS}/\Delta V_{CB}$	0 ≤ V_{CB} ≤ V_{MAX}, (Note 7) 1μA ≤ I_C ≤ 1mA	—	10	25	—	10	50	μV
Offset Voltage Change vs Collector Current	$\Delta V_{OS}/\Delta I_C$	V_{CB} = 0V 1μA ≤ I_C ≤ 1mA	—	5	25	—	5	50	μV
Offset Current Change vs V_{CB}	$\Delta I_{OS}/\Delta V_{CB}$	0 ≤ V_{CB} ≤ V_{MAX}	—	30	70	—	30	70	pA/V
Average Offset Voltage Drift	TCV_{OS}	10μA ≤ I_C ≤ 1mA, 0 ≤ V_{CB} ≤ V_{MAX}, (Note 4) V_{OS} Trimmed to Zero, (Note 3)	— —	0.08 0.03	0.3 0.1	— —	0.08 0.03	1 0.3	μV/°C
Bulk Resistance	r_{BE}	1μA ≤ I_C ≤ 10mA	—	0.3	0.5	—	0.3	0.5	Ω
Collector-Base Leakage Current	I_{CBO}	V_{CB} = V_{MAX}	—	25	200	—	25	400	pA
Collector-Emitter Leakage Current	I_{CC}	V_{CE} = V_{MAX} (Notes 4, 6)	—	35	200	—	35	400	pA
Collector-Emitter Leakage Current	I_{CES}	V_{CE} = V_{MAX} (Notes 4, 6)	—	35	200	—	35	400	pA
Noise Voltage Density	e_n	I_C = 1mA, V_{CB} = 0 fo = 10Hz fo = 100Hz fo = 1kHz fo = 10kHz	— — — —	1.6 0.9 0.85 0.85	2 1 1 1	— — — —	1.6 0.9 0.85 0.85	3 2 2 2	nV/√Hz
Collector Saturation Voltage	$V_{CE\ SAT}$	I_C = 1mA I_B = 100μA	—	0.05	0.1	—	0.05	0.2	V
Input Bias Current	I_B	I_C = 10μA	—	—	25	—	—	34	nA
Input Offset Current	I_{OS}	I_C = 10μA	—	—	0.6	—	—	1.3	nA
Input Offset Current Drift	TCI_{OS}	I_C = 10μA (Note 4)	—	40	90	—	40	150	pA/°C
Breakdown Voltage	BV_{CEO}		40	—	—	40	—	—	V
Gain-Bandwidth Product	f_T	I_C = 10mA, V_{CE} = 10V	—	200	—	—	200	—	MHz
Output Capacitance	C_{OB}	V_{CE} = 15V, I_E = 0	—	23	—	—	23	—	pF
Collector-Collector Capacitance	C_{CC}	V_{CC} = 0	—	35	—	—	35	—	pF

Notes:

1 Current gain is measured with Collector-Base Voltage (V_{CB}) swept from 0 to V_{MAX} at the indicated collector currents.

2 Current Gain Match (Δh_{FE}) is defined as:

$$\Delta h_{FE} = \frac{100 \cdot \Delta I_B \cdot h_{FE}min}{I_C}$$

3 The initial zero offset voltage is established by adjusting the ratio of I_{C1} to I_{C2} at T_A = 25°C. This ratio must be held to 0.003% over the entire temperature range. Measurements are taken at the temperature extremes and 25°C.

4 Guaranteed by design.

5 Sample tested.

6 I_{CC} and I_{CES} are verfied by measurement of I_{CBO}.

7 This is the maximum change in V_{OS} as V_{CO} is swept from 0V to 40V.

Table 2.

$V_S = \pm 15V$, $T_A = 25°C$, unless otherwise noted.

PARAMETER	SYMBOL	CONDITIONS	OP-27A/E MIN	OP-27A/E TYP	OP-27A/E MAX	OP-27B/F MIN	OP-27B/F TYP	OP-27B/F MAX	OP-27C/G MIN	OP-27C/G TYP	OP-27C/G MAX	UNITS
Input Offset Voltage	V_{OS}	(Note 1)	—	10	25	—	20	60	—	30	100	µV
Long-Term Vos Stability	Vos/Time	(Note 2)	—	0.2	1.0	—	0.3	1.5	—	0.4	2.0	µV/Mo
Input Offset Current	I_{OS}		—	7	35	—	9	50	—	12	75	nA
Input Bias Current	I_B		—	±10	±40	—	±12	±55	—	±15	±80	nA
Input Noise Voltage	e_{np-p}	0.1Hz to 10Hz (Notes 3,5)	—	0.08	0.18	—	0.08	0.18	—	0.09	0.25	µVp-p
Input Noise Voltage Density	e_n	fo = 10Hz (Note 3) fo = 30Hz (Note 3) fo = 1000Hz (Note 3)	— — —	3.5 3.1 3.0	5.5 4.5 3.8	— — —	3.5 3.1 3.0	5.5 4.5 3.8	— — —	3.8 3.3 3.2	8.0 5.6 4.5	nV/√Hz
Input Noise Current Density	i_n	fo = 10Hz (Notes 3,6) fo = 30Hz (Notes 3,6) fo = 1000Hz (Notes 3,6)	— — —	1.7 1.0 0.4	4.0 2.3 0.6	— — —	1.7 1.0 0.4	4.0 2.3 0.6	— — —	1.7 1.0 0.4	— — 0.6	pA/√Hz
Input Resistance — Differential-Mode	R_{IN}	(Note 4)	1.5	6	—	1.2	5	—	0.8	4	—	MΩ
Input Resistance — Common-Mode	R_{IN}	(Note 4)	1.5	6	—	1.2	5	—	0.8	4	—	MΩ
Input Voltage Range	IVR		±11.0	±12.3	—	±11.0	±12.3	—	±11.0	±12.3	—	V
Common-Mode Rejection Ratio	CMRR	$V_{CM} = \pm 11V$	114	126	—	106	123	—	100	120	—	dB
Power Supply Rejection Ratio	PSRR	$V_S = \pm 4V$ to $\pm 18V$	—	1	10	—	1	10	—	2	20	µV/V
Large-Signal Voltage Gain	A_{VO}	$R_L \geqq 2k\Omega$, $V_O = \pm 10V$ $R_L \geqq 600\Omega$, $V_O = \pm 10V$	1000 800	1800 1500	— —	1000 800	1800 1500	— —	700 600	1500 1500	— —	V/mV
Output Voltage Swing	V_O	$R_L \geqq 2k\Omega$ $R_L \geqq 600\Omega$	±12.0 ±10.0	±13.8 ±11.5	— —	±12.0 ±10.0	±13.8 ±11.5	— —	±11.5 ±10.0	±13.5 ±11.5	— —	V
Slew Rate	SR	$R_L \geqq 2k\Omega$ (Note 4)	1.7	2.8	—	1.7	2.8	—	1.7	2.8	—	V/µs
Gain Bandwidth Prod.	GBW	(Note 4)	5.0	8.0	—	5.0	8.0	—	5.0	8.0	—	MHz
Open-Loop Output Resistance	R_O	$V_O = 0$, $I_O = 0$	—	70	—	—	70	—	—	70	—	Ω
Power Consumption	P_d	V_O	—	90	140	—	90	140	—	100	170	mW
Offset Adjustment Range		$R_p = 10k\Omega$	—	±4.0	—	—	±4.0	—	—	±4.0	—	mV

Notes:

1. Input offset voltage measurements are performed — 0.5 seconds after application of power. A/E grades guaranteed fully warmed-up.
2. Long-term input offset voltage stability refers to the average trend line of Vos vs. Time over extended periods after the first 30 days of operation. Excluding the initial hour of operation, changes in Vos during the first 30 days are typically 2.5µV — refer to typical performance curve.
3. Sample tested.
4. Guaranteed by design.
5. See test circuit and frequency response curve for 0.1Hz to 10Hz tester.
6. See test circuit for current noise measurement.

Table 2. Main electrical characteristics of the OP-27 operational amplifier.

parallel-connected 1% resistors: this is, strictly speaking, not necessary, but is done to enable constructors making up the exact values with other combinations of resistors.

Capacitors C_{12} and C_{13} limit the DC amplification of the opamp to unity. Automatic offset correction in this stage was decided against because of the requirement to suppress very-low-frequency (below 5 Hz) components. Although the nominal gain of IC_2 is only 14 dB, the frequency-selective networks cause an additional gain of 20 dB for signals below 50 Hz.

The coupling capacitors in the output circuit of IC_2 are not strictly necessary, because the automatic offset correction at the input stages works so well that there is no discernible DC at the output of IC_2.

The supply lines to the different stages are decoupled separately. The relevant network consists in each case of two 1000 µF electrolytic capacitors, each shunted by a 220 n ceramic capacitor for better high-frequency operation. Each electrolytic capacitor is connected in series with a low-value resistor to further improve the decoupling.

Line amplifier

The quality of the line amplifier is particularly important for the faithful reproduction of compact discs, when a good dynamic range, broad bandwidth, and minimal distortion are essential.

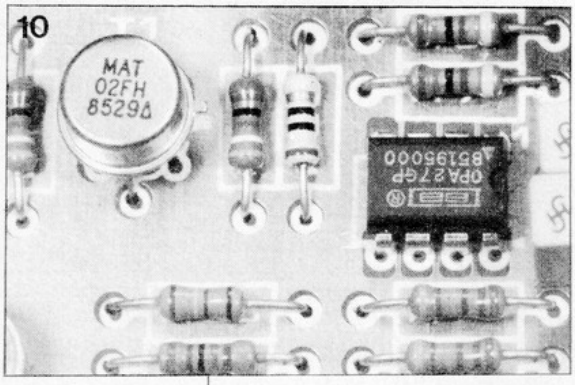


Fig. 10. The two important semiconductor devices used in the preamplifier: dual transistor MAT-02 and operational amplifier OP-27.

The dynamic range is determined by the maximum supply voltage to the ICs, which here has been made as high as possible at ±18.5 V. This value makes possible an undistorted output voltage of about 12 V, which is ten times as high as the nominal output level of 1.2 V (to give about 20 dB headroom).

Since the noise produced by opamps is very low, and the gain of the devices is only about 14 dB, the signal-to-noise ratio is of the order of 100 dB. With the headroom of about 20 dB, this gives a total dynamic range of around 120 dB!

As stated in Part 1, the voltage dividers at the inputs merely serve to reduce crosstalk and hardly attenuate the wanted signal. Only the CD input is provided with an attenuator to lower the signal level by about a half. The reason for this is that the majority of CD players provide a fairly high ouput signal—of the order of 1 V. The closer the CD player ouput is to the nominal sensitivity of the preamplifier, the smaller the likelyhood of overload during peak levels. The signal-to-noise ratio is not affected in the least by the attenuator (bear in mind that the output level control on the CD player performs the same function).

The circuit of the line amplifier—see Fig. 11—consists of two Types OP-27 opamps per channel. The two-stage arrangement has the benefit of a greater dynamic drive range, because, since the volume control is normally nowhere near fully open, and, like the balance control, is fitted between the two amplifiers, the first opamp can deliver a greater undistorted signal without overloading the second opamp. This set-up has the further advantage that both controls are isolated from the inputs and the output.

The first stage, IC_4 (IC_4'), has a gain of 6 dB. Its output is connected to the second amplifier via a stereo/mono switch (and, as already stated, the volume and balance controls). The switch is actually a relay contact to obviate long signal paths to a conventional switch. The 1K resistor, R_{50} (R_{50}'), ensures that the opamps do not short-circuit each other's output in the mono condition.

Each channel has its own individual (mono) balance control, P_1 (P_1'), but the volume control, P_2 , is of the customary stereo type. The individual balance controls enable setting the output signal to maximum level. Moreover, good-quality stereo balance controls are virtually unobtainable!

The second stage, IC_5 (IC_5'), has a gain of about 10 dB, resulting in the line amplifier delivering an output signal of 1.2 V for an input of 200 mV. The output is provided with DC blocking capacitors: again, these are not strictly necessary, since the opamps have no offset problems. But it could just happen that one of the signal sources delivers a DC component, and this would be amplified together with the signal.

The output terminals are connected to the opamps via relay contacts: the relay action is delayed at switch-on, but is immediate on switch-off. The output terminals are also disconnected briefly when the inputs are switched to obviate annoying clicks.

As in the phono amplifier, the supply lines are thoroughly decoupled. Each opamp is supplied via a separate 10R resistor, and individually decoupled by two 1000 µF electrolytic capacitors, each of which is shunted by a 220 n ceramic capacitor for improved high-frequency operation.

As will be seen next month in the constructional details, the earth tracks of the two channels are kept separate on the printed-circuit board, and combined only at the main earth rail. This arrangement improves the already excellent channel separation figures.

11

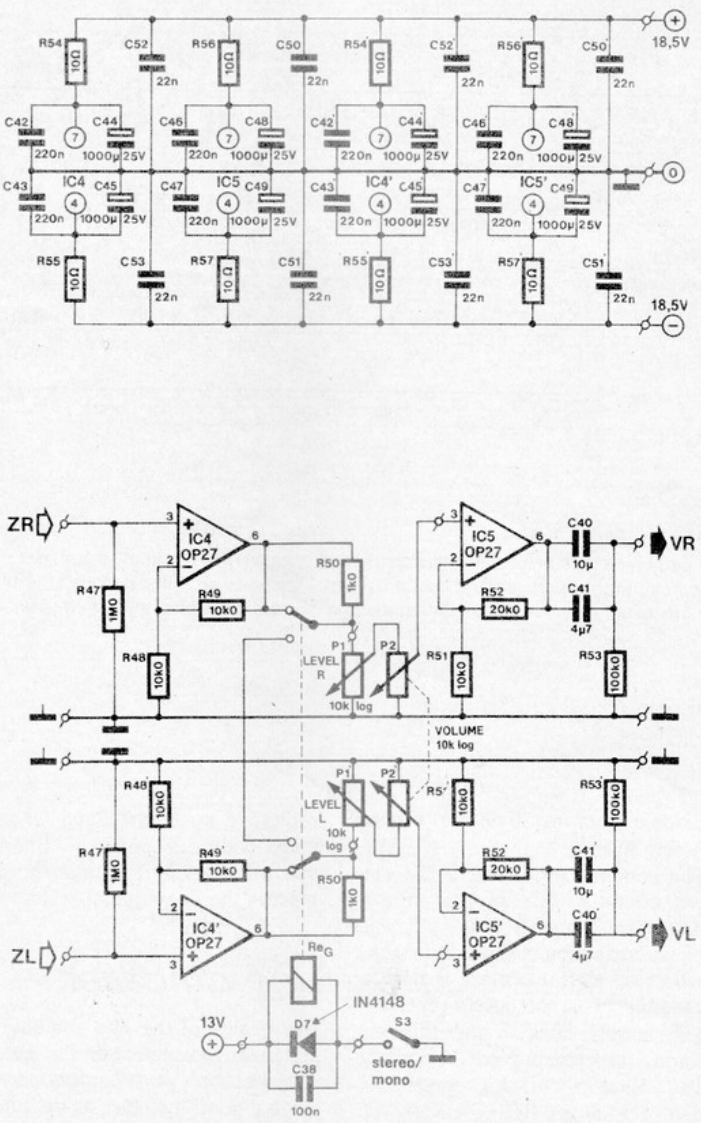


Fig. 11. The circuit diagram of the line amplifier. This amplifier is of particular importance in the reproduction of sounds from a compact-disc player. The controls have been inserted between the two opamps contained in the amplifier to prevent them having any effect on the input or output impedances.

ATLAS, MINERVA, CARMINAT, CARIN AND PROMETHEUS

Not Greek gods but Renault's electronic cars of the future

From the closely-guarded research laboratories of Renault, first details have now been released on some of the French company's spectacular advances into the cars of tomorrow's world, where computers and microchip electronics will take over the "thinking" aspects of motoring — from actual navigation and route-finding to the timing of departure and arrival, selecting alternative routes to avoid traffic jams, and even pin-pointing the car's location at any given time.

All this has evolved into a language of its own, embracing words like ATLAS, CARMINAT, CARIN and PROMETHEUS — acronyms for the various research projects now in hand.

The following feature provides a fascinating insight into the current and future stages of "robotic" cars and motoring being developed by Renault and its European associates.

Project ATLAS

In today's world, communication and information are growing in importance day by day. This tendency is reflected in the development of the automobile by increased integration of electronics in the passenger compartment — an already apparent and indisputable evolution.

The motorist now regards his vehicle less as "wheels" and more as an integral part of his environment. Information for road users is becoming increasingly organized with the creation of radio-guides, organized planning of peak holiday departures, and the setting-up of information centres.

As a result, the motorist wants more information — and to be able to communicate, not just with his vehicle but with the outside world.

It is against this background that in 1981 Renault launched a research programme whose objective was to obtain:

— a device integrated into the car enabling selected supplementary information to be given to the driver, and

— an evaluation "tool" to define new automobile equipment, associated with the following:—

* The development of new electronic technology (sensors, capacities for handling information, miniaturisation, etc.
* The services offered by organizations outside the car.
* Improvements in reliability, through a reduction in the number of components and controls used by the driver.

Project ATLAS (Acquisition through Telediffusion of Automobile Logistics for Services) is the result of this research, undertaken with Telediffusion of France (TDF) since 1982.

ATLAS makes it possible to treat in real time and display on an interactive, touch-sensitive screen, information for the driver which can be classified under three categories:—

Information related to the vehicle itself, said to be endogenous, and generated by sensors (warnings, maintenance, diagnostic, and general mechanical condition). Importantly, only information made compulsory by regulations features on the dashboard (speed, fuel level, mileage, etc.)—all other information being available on the screen.

This redistribution of the siting of information is aimed to increase the safety of the "transport" function and, at the same time, increase the amount of information available.

Pre-recorded information, described as "loaded", and available on a compact disc, or memory card (for example, the Renault dealer network list, a practical guide to Renault equipment, extracts from the car handbook, perusal of microprocessor card, etc.).

Information outside the vehicle, categorised as "exogenous" which will be transmitted by a telediffusion network (road information, traffic, alternative routes, weather forecasts, etc.). As the amount of information proposed could be very large, it is updated and transmitted in real time.

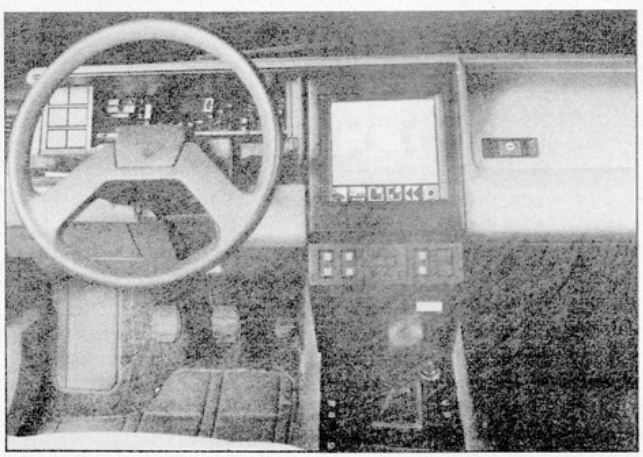
Project ATLAS makes current evolution concrete fact, through:

— appearance of new banks of facts;

— development of the automobile product;

— new expectations of motorists in matters of comfort and functional dependability, and

— changes in the behaviour of users, allied to the scope of communications and the evolution of social relations.



Development of the ATLAS system

The first stage of research work made it possible to present Renault's "DIALOG" (voice response) model at the Paris Motor Show in October, 1984. A road-going system on a Renault Espace 2000 was also presented by TDF at the Montreux Symposium in June, 1985.

Analysis of the first elements obtained led Renault to centre its

research work on the development and treatment of "endogenous" information and diagnostics (collection of information, treatment and design, ergonomics and integration into the passenger compartment). These aspects are in fact more within the domain of the car maker.
All matters relating to the "exogenous" part of Project ATLAS were entrusted to the Société SAGEM (Société d'Applications Générales d'Electricité et de Méchanique) because of its competence in on-board electronics, synthetic images and artificial intelligence, and navigation.
TDF will continue its work on the transmission of information because of its specific competence in this field. Contact was also made with Philips in connection with a numerical information receptor.
"Finally", says Renault, "it should be remembered that systems of the ATLAS type have European vocation, both at the level of service (road information, and more generally the whole of the information given), and at the technical level (European minimum standards).
"Also, among the technologies developed within the framework of the ATLAS system, most seem to conform with the declaration of principle adopted by the European Ministers at the second ministerial conference on the EUREKA project at Hanover on November 5/6, 1985:
— transport technology;
— information and communication;
— protection of the environment.
"Thus, the ATLAS system will only find its full dimension on a European scale. In the same way, it favours normalisation of transport and exchange within the European Community."
As a result of these developments, Renault put in hand a dossier within the framework of the EUREKA project, christened MINERVE (Media Intelligent pour l'Environnement Routier du Véhicule Européen), at the same time continuing its own work. Control of the MINERVE project was entrusted to Société SAGEM.

Project CARMINAT

More recently, Philips has been working in parallel with ATLAS on the CARIN project (Car Information and Navigation). This project consists of fitting vehicles with an **electronic co-driver**, able to:
— determine the itinerary;
— guide the driver to his destination;
— give the position of the vehicle, and indicate it at any time,
— and to supply information on the environment or the destination.
In the first instance, particular effort was devoted to navigation and the use of a compact disc in the vehicle. The CARIN system was first demonstrated in 1985. Philips, contacted by SAGEM in relation to the MINERVE dossier, wished like Renault to resolve this problem on a European scale.
The companies decided to pool their efforts and experience in a common project — CARMINAT. This combines the knowledge acquired through CARIN, MINERVE and ATLAS, and was presented within the framework of the EUREKA project.
It should lead, in 1989/90, to a range of products which can be used on European vehicles.

Project PROMETHEUS

On the initiative of Daimler-Benz, and with the active support of Renault in view of its experience with the ATLAS project, European automobile constructors have developed an extensive research programme whose objective is to create concepts and solutions for fluid traffic movements, with reduced impact on the environment and increased economy, combined with maximum safety.
This vast programme broadens and extends the actions already described. It has been baptised "PROMETHEUS", acronym for Programme for a European Traffic with Highest Efficiency and Unprecedented Safety, and falls perfectly within the framework of the dynamics of the European programme EUREKA.
Project PROMETHEUS also brings in the fields of fundamental research essential to attain the objectives set (micro-electronics, expert systems and artificial intelligence, and communication diffusion technology).
Many European experts are associated in the project. This is how the following French institutes are associated with PROMETHEUS as experts:
— CNRS, principally implicated in the scientific knowledge of artificial intelligence problems and advanced electronics;
— INRIA, specialists in advanced techniques of artificial vision and identification of forms and movements, together with the treatment of speech;
— INRETS, for traffic questions and dynamic information;
— CCETT of Rennes (Centre Commun d'Etudes de Télédiffusion et de Télécommunications), specialists in these areas.
Automobile equipment suppliers and industrialists in electronics will be associated with the project after the one-year preliminary phase to define the technological specifications resulting from the following three axes of research:—
PRO-CAR: development of a computer "co-driver", to improve vehicle safety;
PRO-NET: development of communications networks from vehicle to vehicle;
PRO-ROAD: development of communication between the road environment and the computer co-driver.
Vehicles will thus receive information making it possible to organize maximum fluidity of traffic.
The specifications contained in ATLAS, MINERVA and finally CARMINAT significantly cover the concepts of PRO-CAR and PRO-ROAD contained in the PROMETHEUS project, and use the same basic research in micro-electronics, expert systems and diffusion techniques.

The major stages of project ATLAS

1981 — Preparation of the ATLAS project briet.
1982 — TDF-RENAULT agreement.
Renault: study of the needs of clients.
TDF: study of the transmission of facts to those on the move.
1983 : Static model.
Renault: verification of functional ergonomics.
TDF: signal reception tests on the road.
— Working model. Construction of a working model fitted with the ATLAS system (Renault 20).
1984 — First reception on board a moving vehicle of a radio-diffused programme (Renault 20).
— Presentation of the DIALOG model at Paris Motor Show (October, 1984).1985-Presentation of the DIALOG model at the Geneva Motor Show (March, 1985).
— Presentation of the ATLAS system in working form on a Renault Espace 2000 at the Montreux Symposium (June, 1985).
1986 — Presentation of the ATLAS system in working form on three vehicles (2 Renault 21 TXEs, 1 Renault Espace 2000).

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TEMPERATURE PROBE FOR DMM

This article deals with a plug-in temperature-to-voltage converter for use with a digital multimeter (DMM). The design includes a battery test facility and a home-made probe for ease of measurement.

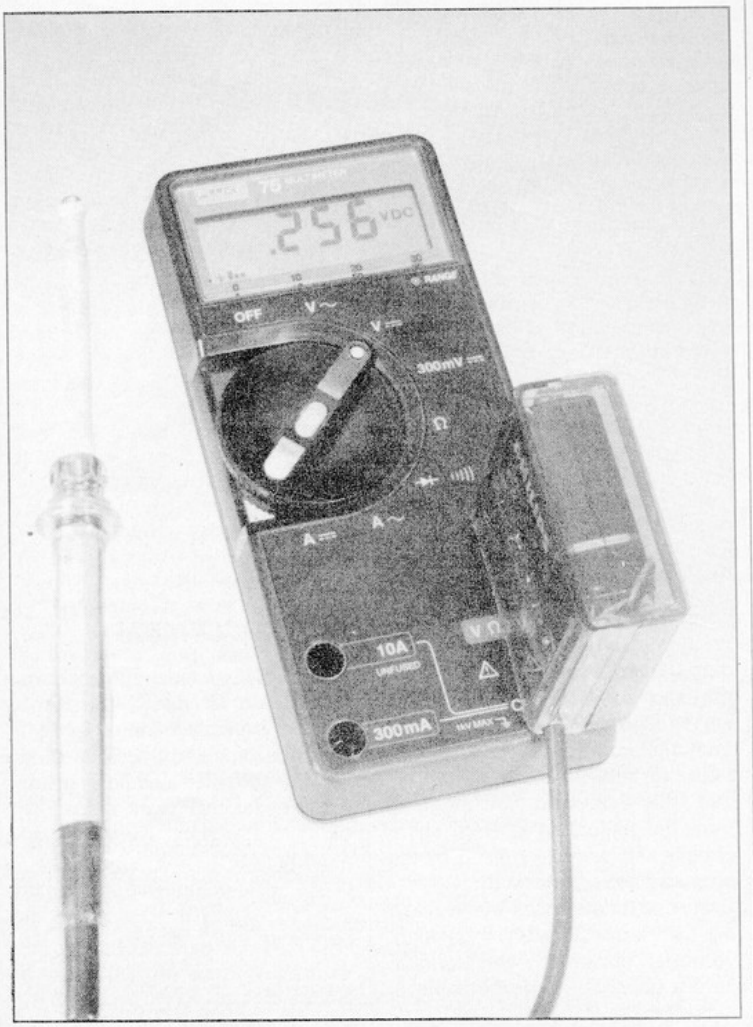
The Type LM35 from National Semiconductor is stated to have the following features:*

- Calibrated direct in degrees Celcius (Centigrade) (°C).
- Linear +10.0 mV/°C scale factor.
- 0.5 °C accuracy at +25 °C (LM35C).
- Rated for full −55 to +150 °C range.
- Suitable for remote applications.
- Operates from 4 to 30 volts.
- Less than 60 μA current drain.
- Low self-heating, 0.8 °C in still air.
- Non-linearity only ±0.25 °C typical.
- Low-impedance output, 0.1 Ω for 1 mA load.

Circuit description

Fig. 1 shows the circuit diagram of the proposed temperature probe. Diodes D_3 and D_4 have been included to obtain a circuit ground potential which is some 1.2 V lower than that of the temperature sensor. Resistor R_5 ensures that the sensor output voltage can be negative with respect to the DMM ground in the case of measuring temperatures below 0 °C. Resistor R_4 decouples the probe output from the high-impedance DMM input (Z_{in} = 1 MΩ, usually).

The remainder of the circuit serves as the sensor supply and the battery test facility. Actuation of push-button S_1 causes the voltage at junction R_2-R_3-C_1 to be nearly equal to V_{batt}, as C_1 is not charged at the onset. Transistor T_1 conducts and D_1 lights if the terminated battery voltage is higher than 7 V. After a predetermined period, C_1 is fully charged, and T_1 turns off the LED. The battery test thus immediately indicates a flat battery if the LED remains off after actuation of S_1.



Construction

A suggested method of constructing the temperature probe is shown in the photograph at the head of this article and in the drawing of Fig. 3. The component overlay and track pattern of miniature PCB Type 86022 is given in Fig. 2.

The temperature sensor proper is conveniently fitted onto the tip of a salvaged, temperature-controlled, soldering iron, whose heating element has been removed from the metal tube.

The LM35 is preferably secured onto the tip with two-component glue, while a 3-wire cable is run through the tube to make the connection to the plug-in unit; the length of the probe cable should be a maximum one metre or so. As shown on the in-

* Data taken from manufacturer's data sheet.

Fig. 1. In essence, the temperature-to-voltage converter is composed of a precision temperature sensor, IC_1, and a battery condition tester.

Fig. 2. Track layout and component mounting plan for the plug-in converter. The small size of the board allows the circuit plus battery to be fitted in a compact enclosure.

Parts list

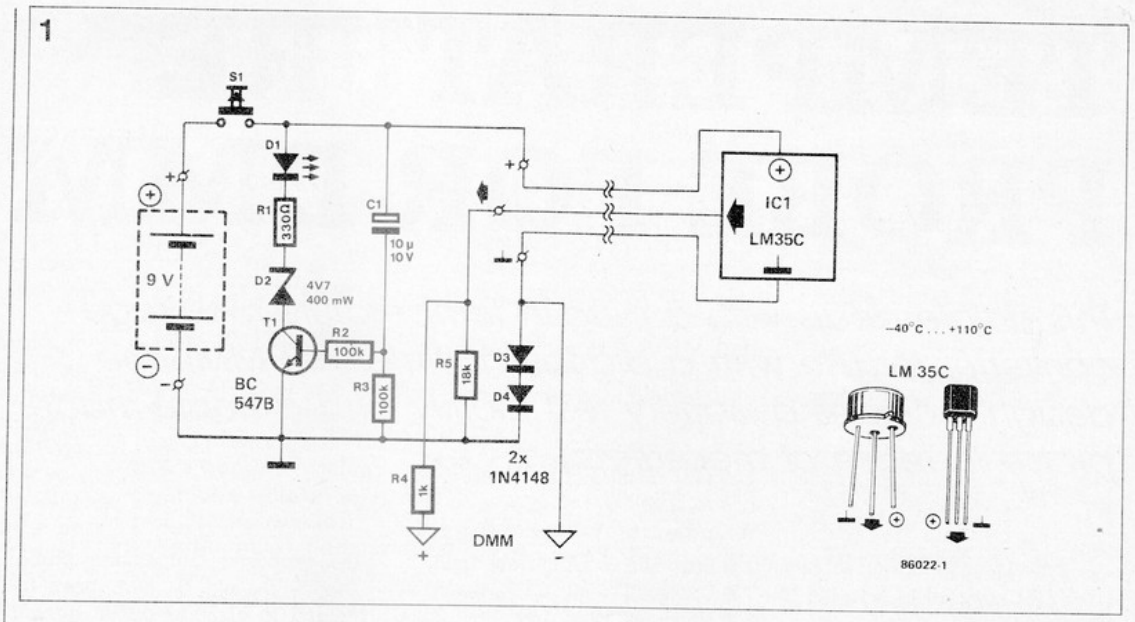
Resistors (1/8 W):
R_1 = 330 Ω
R_2;R_3 = 100 k
R_4 = 1 k
R_5 = 18 k

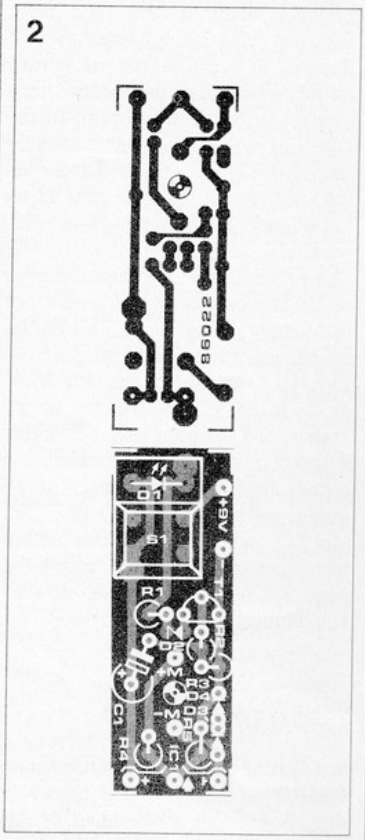
Capacitor:
C_1 = 10 µ;10 V

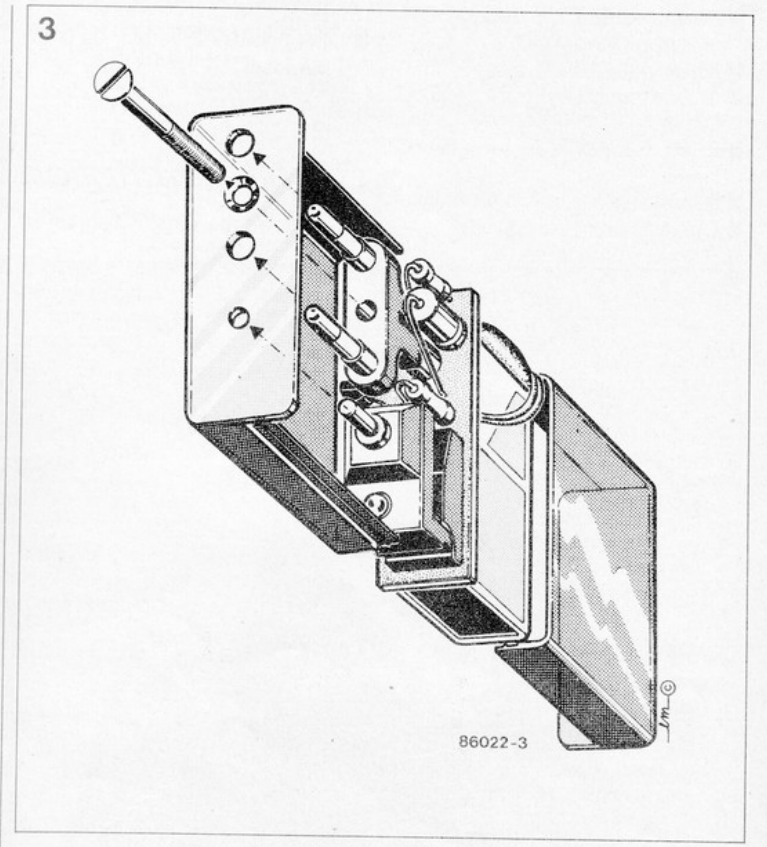
Semiconductors:
D_1 = LED (may be mounted in Digitast keytop)
D_2 = 4V7 zenerdiode 0.4 W
D_3;D_4 = 1N4148
T_1 = BC547B
IC_1 = LM35C

Miscellaneous:
S_1 = ITT Digitast Type S switch (illumination is optional)*
2 off wander plugs *
PP3 battery plus clip
Probe tube *
PCB type 86022 (see Readers Services)

* See text







troductory photograph, the probe tip may be insulated with a short length of heat shrink sleeving. If this is done, however, the sensor enclosure should remain uninsulated.

Finally, the tube is carefully sealed to enable measuring the temperature of liquids.

As shown in Fig. 3, the completed board and PP3 battery are housed in a transparent plastic case. Although a Euro-type, moulded mains plug makes for a very simple connection to the DMM input sockets by virtue of the correct pitch, two correctly spaced wander type plugs will also do quite nicely.

As S_1 is to be actuated when plugging the unit onto the DMM, a small pin must be fitted onto the Digitast keytop. This pin is made to protrude from the converter enclosure and tested for reliable action.

The PCB should be well insulated from the battery to prevent short-circuits and damage caused by the corrosive battery contents.

Every constructor is left free to make his own, approximately 5 cm high, converter enclosure, which should have a hole for the probe cable to enter.

Fig. 3. Artist's impression of a suggested direct plug-on enclosure: S_1 is fitted with a protruding pin to achieve instantaneous battery test and converter operation as the unit is plugged onto the DMM DC input sockets.

Applications

As the LM35 provides a linear output of +10 mV/°C, the DMM display reading is simply the measured temperature, provided you have grown accustomed after a while to imagining the decimal point shifted two digits to the right. For example, a reading of 0.256 V represents a probe temperature of 256 mV/10 = 25.6 °C. Similarly, —0.307 V re--peresents —30.7 °C.

Th

MW receiver

a straightforward 'straight-through' receiver

Why should there be a need to construct your own medium waveband receiver? Surely it is far cheaper to buy one at the local supermarket? This may well indeed be true, but it is far more fun to actually build one yourself. After all, many of our readers belong to the younger generation and there is nothing quite like building your first radio — and getting it to work! — as many of our more experienced readers will testify.

Amateur constructors often feel like magicians. It is quite amazing what can be accomplished with very few components. Take the design for this receiver for instance; an RF amplifier and a couple of transistors to bring music to your ears! In any case, many readers felt that it was high time that a simple receiver circuit found its way into the EPS list once more.

The object of the exercise is to end up with a neat, economical portable radio. One that fits comfortably inside a coat pocket and can keep you up to date with the latest news and pop music, as you travel around town. Another important factor, of course, is that a single 9 V battery should last as long as possible (a few months at least).

When designing such a project, the first choice has to be between AM and FM. Nowadays, FM is favourite, but the problem here is that is not so easy for the novice to build, especially if the finished unit is to be really small.

Elektor does have printed circuit boards available for something a little larger than that described here, but by no means one that requires so few components, it can be virtually put together with your eyes closed. Which is our main objective, remember.

We therefore came to the conclusion that there is nothing wrong with the medium waveband. It certainly has not run out of stations yet and what is more, the set will be much simpler (and cheaper) to build than an FM radio. It can be far smaller in size and last, but by no means least, it needs no finnicky aerial. In other words, it really is a pocket radio.

Superhet or superreg?

Now that we have decided upon medium wave and the main requirements are that it be small, simple to build and conservative on batteries, we need to work out a few more design parameters.

The majority of manufactured radio receivers operate on the superheterodyne principle. However, most single waveband receivers utilise the superregenerative principle. This is, in fact, *the* recipe for a reliable receiver if it is to have a fairly high performance and feature reasonable sensitivity in spite of its compact size. Nevertheless, if such considerations as simplicity of construction and ease of calibration are involved, the 'super' part is best omitted. This is further illustrated in figure 1. All the most common AM receiver principles are shown there.

First the 'straight-through' receiver (a). This is comprised of an adjustable LC tuned circuit, a high frequency amplifier, a detector, an audio amplifier, a detector, an audio amplifier and a loudspeaker. The RF stage could even be left out, so that the set would then be a 'sophisticated' crystal receiver. If it is to be sufficiently sensitive, however, rather a lot of RF amplification will be necessary. This is why the RF amplifier usually incorporates an adjustable feedback network (see dotted line) which enables the set to be adjusted to the point of oscillation (maximum sensitivity) for every station.

The reflex receiver in figure 1b also offers a reasonable degree of sensitivity. Here the RF amplifier stage is not only used in the conventional manner, but it also amplifies the audio signal. This type of receiver used to be very popular in the days when transistors were rather expensive and difficult to obtain.

Figure 1c shows that even the 'simple' superhet can be quite complicated. The aerial signal is now 'added' to that of an oscillator in the mixing stage. The oscillator produces a somewhat higher (or lower) frequency than the input signal and is varied simultaneously with the tuning capacitor. This generates a constant 'sum' or 'difference' frequency regardless of the actual frequency of the input signal. This 'intermediate' (IF) frequency is filtered at the output of the mixer and is further amplified. If necessary, the signal can be filtered and amplified several times to improve the selectivity. This is because the constant frequency of the IF signal makes the tuning of the LC circuits for each station superfluous. Obviously, the receiver will be rather complicated to set up.

Straight-through

Seeing that practically all the receivers that have been published in Elektor over the past few years were either superheterodyne or superregenerative, our design staff thought that it was time that a simple version was produced. In any event, an IC exists which will fit the bill perfectly, but more about this later. Thus, after due consideration, the recipe illustrated in figure 1a was chosen for the miniature MW receiver, albeit without the feedback stage. The latter, even in the version shown in figure 1b, will make any receiver a lot less portable. Also, construction becomes a critical task, the set is likely to 'whistle' and more often than not the receiver will have to be operated with both hands as the amount of feedback has to be adjusted for each individual station. If an ordinary 'straight-through' receiver can be built to incorporate enough RF amplification for feedback to become superfluous, without causing it to oscillate, it will have many practical advantages. The miniature integrated circuit that we have in mind does just this and furthermore features other useful characteristics, as will be seen later.

Compared to more usual sets, a simple single tuned circuit receiver (such as this one) will be much less selective and therefore not so sensitive. Since MW receivers, especially pocket-sized ones, are more often than not used for reception of a limited number of local radio stations, this disadvantage will not be so noticeable. It is amply compensated by the following advantages over other types:

- it is much easier to build
- it does not require any alignment
- it does not include an oscillator, thereby avoiding stability problems
- no mixing is involved, reducing 'whistle' considerably

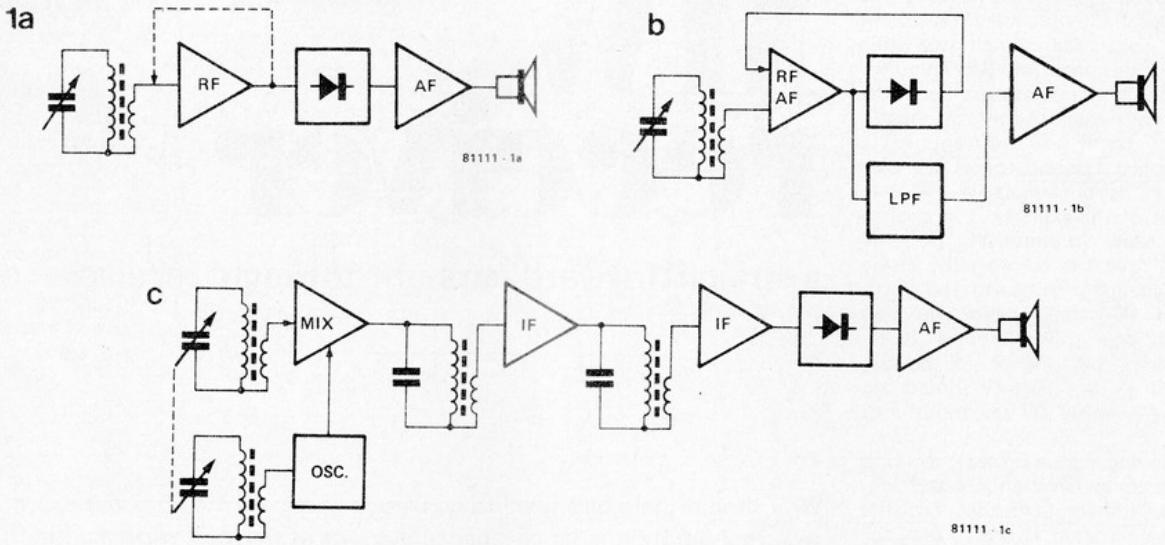


Figure 1. The three most common types of AM receiver: a straight-through receiver (a), a reflex receiver (b) and a superheterodyne receiver (c).

- its sound is of superior quality to that of the average superhet

The ZN 414

By far the easiest method of constructing a straight-through medium, waveband receiver is to use the ZN 414 integrated circuit from Ferranti which was designed specifically for this purpose.

Having only three pins, it looks more like a transistor than a 'proper' IC. Although it has been around for quite some time, it continues to provide the best solution for a receiver where a minimum number of components is required. This is clearly illustrated in the diagram in figure 2. It shows the complete MW receiver constructed around the ZN 414. All that is required is a single transistor amplifier stage to provide a first class matchbox receiver.

Certain items immediately catch the eye. First, the low supply voltage. The ZN 414 is designed to be powered from a single battery. Its supply voltage range is between 1.2 and 1.6 V and the current consumption is in order of 0.3 mA. This device could hardly be more economical.

Also remarkable is the fact that the coil (L1) consists of a single winding, instead of the usual double-wound or tapped inductor, and that the detector diode which one would usually expect is missing.

The double wound coil is superfluous as the IC features an extremely high input impedance (4 MΩ) which is only a very slight load for the parallel tuned circuit. Not only does this make the coil that much easier to wind, but also it helps to prevent interference from short wave transmitters. As far as the detector diode is concerned, this is already integrated in the IC in the form of a transistor detector which uses capacitor C3 as the only external component.

2

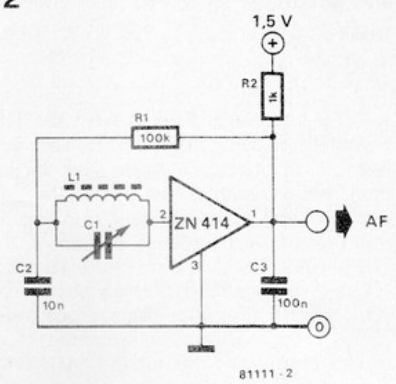


Figure 2. What could be simpler? The straight-through receiver using the ZN 414 IC.

The block diagram of the medium waveband receiver, see figure 3, shows just what goes on inside the case of the ZN 414 (see inside the dotted area). It is comprised of a high impedance input stage (drawn here as an emitter follower), a (three stage) RF amplifier with a frequency range of 150 kHz . . . 3 MHz and a gain of 72 dB, an AM detector and, finally, an automatic gain control (AGC).

Too much should not be expected from the latter, as its range is about 20 dB, just enough to smooth out any slight differences in amplitude between the various radio stations. As soon as the unit is in close proximity to a powerful transmitter, however, the automatic gain control will be unable to adjust to the particular station required. Nevertheless, it is far better to have 20 dB than none at all, as is the case in some elementary receivers.

Circuit diagram

From the block diagram in figure 3 it can be seen how straightforward the complete pocket sized medium wave receiver is. Apart from the parallel tuned circuit, the ZN 414 and the audio amplifier, all it needs is a suitable circuit to derive the 1.3 V required by the ZN 414 from the power supply for the audio amplifier. A simple bleed

3

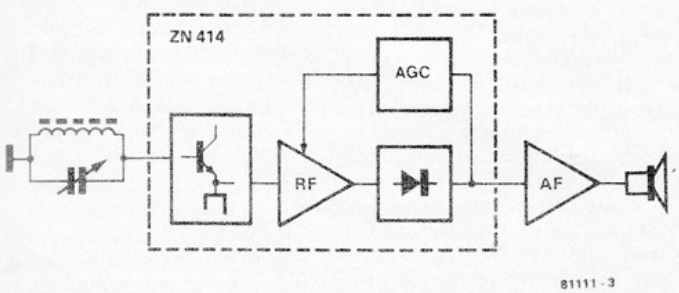


Figure 3. The block diagram of the miniature MW receiver. The entire section inside the dotted area is incorporated inside a single IC no bigger than a BC 107.

4

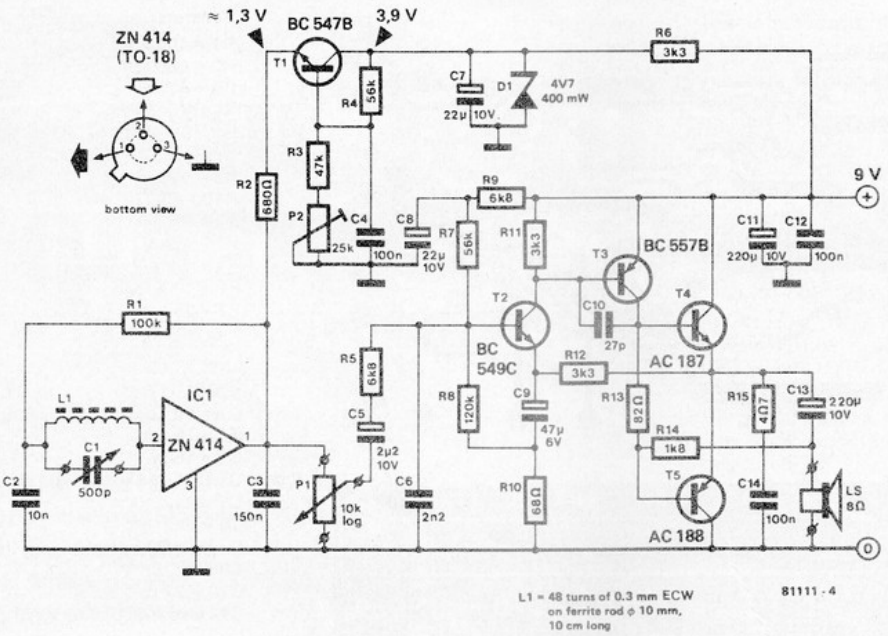


Figure 4. The circuit diagram of the complete MW receiver featuring reasonable sensitivity, good selectivity and good quality sound.

resistor and zener diode would have been more than adequate, but a far better method has been employed here.
The complete circuit diagram of the MW receiver is shown in figure 4. The actual receiver section is constituted by IC1 and the surrounding components and is, of course, identical to the diagram shown in figure 2. The only difference between the two is that the values of R2 and C3 have been slightly modified. This is because they are based on the ideal supply voltage for the ZN 414, which is between 1.3 and 1.4 V.

When calculating the value of R2 three parameters have to be taken into account. First, the ratio R1/R2 will affect the automatic gain control. Since the value of R1 must be 100 kΩ, only the value of R2 can be altered. Therefore, the value of the latter will also affect the gain of the ZN 414 and, as the voltage supply to the IC must be at a constant level, the gain will be reduced if a relatively high value is chosen for R2. Moreover, it is important that the values of R2 and C3 constitute a low pass filter with a turnover frequency of around 4 kHz, which is necessary for the detector included in the IC.

The solution, therefore, is to select the best compromise value for R2 and to find an effective method of regulating the supply voltage for the ZN 414. This is why the voltage source constructed around transistor T1 has been added to the circuit. The voltage on the emitter of T1 can be adjusted between approximately 1.2 and 1.45 V by means of the preset potentiometer P2. This may not seem to be very much, but it affects the gain of the IC somewhat considerably. This is an advantage as the sensitivity of the receiver can now be adapted to specific circumstances by presetting the gain of the amplifier as required. Obviously, this will be at a maximum in isolated areas and lower in the vicinity of powerful local transmitters so that the set is not overdriven, which could cause distortion and poor selectivity.
Batteries spend a very brief period of their lifespan at their rated nominal voltage and for this reason, together with the fact that the supply voltage for IC1 is critical, the voltage source T1 is not fed directly from the battery. Instead, the voltage is first regulated by a zener diode (D1) to iron out any fluctuations in the battery voltage. Since the receiver also has to be economical, the current supply to the zener diode is limited by means of a fairly large series resistor (R6). As the current consumption of the ZN 414 is very low the zener diode will operate very well, even at a lower voltage than that which it is rated at (about 3.9 V in this case).
So much for the receiver section. Now for the audio amplifier. Initially, it was proposed that one of the well-known amplifier ICs should be used. These, however, turned out to rapidly exhaust the small 9 V battery's current supply. Instead, it was decided to combine two pnp transistors and two npn transistors to form a discrete amplifier. Very little can be said about this, as it is constructed entirely according to the 'four transistor recipe'. It requires very little current and there is virtually no quiescent current for the output transistors T4 and T5. Thus, when there is no input signal the entire amplifier will only consume around 2.5 mA.
Since the current requirement for the ZN 414 is also fairly modest, the receiver will only consume a total of 4 mA. A reasonable battery will therefore last a considerable time, provided the volume is not turned up to an ear-splitting level, that is.
The maximum output power of the audio amplifier is in the region of 250 mW. In theory it will produce more from a 9 V battery (about 1 W maximum into 8 Ω), but the voltage gain of the audio stage is limited so that no more than 4 V_{pp} is available across the loudspeaker output even when the output signal from the ZN 414 is at a maximum (approximately 30 mV_{eff}). This maintains the current consumption at a level acceptable to the 9 V battery and also eliminates the need for heatsinks on the two output transistors.

Construction

The printed circuit board and component overlay for the medium waveband receiver are shown in figure 5. The only components which are not actually mounted on the board are the variable capacitor C1, potentiometer P1 and the

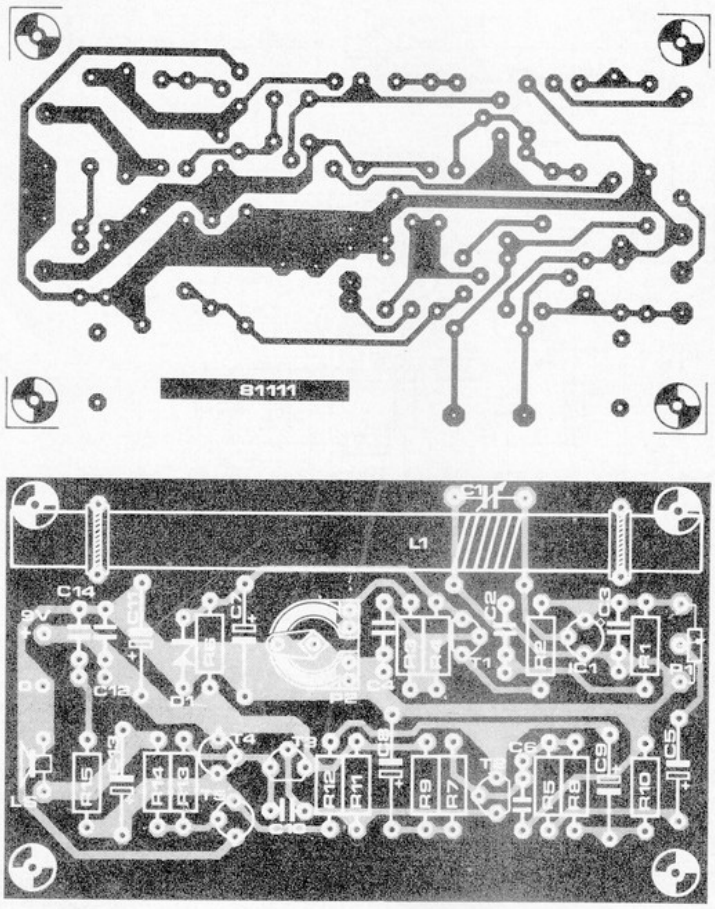


Figure 5. The printed circuit board and component overlay for the MW receiver.

Parts list

Resistors:
R1 = 100 k
R2 = 680 Ω
R3 = 47 k
R4,R7 = 56 k
R5,R9 = 6k8
R6,R1,R12 = 3k3
R8 = 120 k
R10 = 68 Ω
R13 = 82 Ω
R14 = 1k8
R15 = 4Ω7
P1 = 10 k log
P2 = 25 k preset

Capacitors:
C1 = 500 p variable capacitor
C2 = 10 n
C3 = 150 n
C4,C12,C14 = 100 n
C5 = 2μ2/10 V
C6 = 2n2
C7,C8 = 22 μ/10 V
C9 = 47 μ/6 V
C10 = 27 p
C11,C13 = 220 μ/10 V

Semiconductors:
T1 = BC 547B
T2 = BC 549C
T3 = BC 557B
T4 = AC 127, AC 187
T5 = AC 128, AC 188
IC1 = ZN 414
D1 = 4V7 zener diode, 250 mW or 400 mW

Miscellaneous:
LS = 8 Ω/0.2 W loudspeaker
L1 = 48 turns of 0.3 mm dia enamelled copper wire on a ferrite rod of 10 mm diameter 10 cm long

loudspeaker. The leads connecting the capacitor to the board should, obviously, be as short as possible.

As you probably already know, the performance of a parallel tuned circuit is largely dependent on the Q of the tuning inductor. For this reason, the aerial coil, L1, will have to be wound with the utmost care and attention. It is best to use the parameters specified, that is, 48 turns of 0.3 mm diameter enamelled copper wire on a ferrite rod with a diameter of 10 mm and a length of 10 cm. The ferrite rod can be mounted onto the board by means of two short pieces of string. Holes have already been drilled in the printed circuit board for this very purpose.

It is a good idea to wind L1 around a paper or cardboard tube so that it can be moved up and down on the ferrite rod later; the permeability of ferrite and ferroxcube material tends to vary, so therefore it may be necessary to 'trim' the receiver if the stations are not at the correct places on the waveband.

Further remarks. Firstly, something that probably does not need mentioning. As the ferrite rod coil is in fact an aerial, it would be unwise to mount the completed receiver in a metal case! Secondly, the zener diode D1 must be either a 250 mW or a 400 mW type, as stated, as otherwise the input level for the voltage source T1 (3.9 V) will not be correct. This is because the current flowing through D1 is far lower than normal in order to keep the current consumption of the circuit to a minimum. Thirdly, as the output transistors do not require any quiescent current, the value of resistors R13 and R14 are fairly critical. If the stated values are not adhered to the chances are that the output transistors will start to draw current after all and, as there is no temperature compensation network, this could well have a detrimental effect on them. Using the values given in figure 4, transistors T4 and T5 will not have to be cooled. They can be ordinary types without any need for heatsinks.

Results

In practice, the miniature medium waveband receiver was found to perform very satisfactorily. Being a single coil type, it may require constant retuning due to the set 'drifting' off frequency, especially where distant stations are concerned. Even so, it is eminently suitable as a 'stand-by' receiver for news bulletins etc. which is quite often all that is required anyway. It is only when the owner wishes to listen to a weak station in the neighbourhood of a powerful one that the MW receiver is going to have problems. This can often be remedied by turning the receiver towards the weaker station thereby eliminating the stronger one. Local stations can be received very well. In unfavourable circumstances, an external aerial may be experimented with. This should be connected to the top of the tuning coil via a small value capacitor (4p7). This, however, should hardly ever be necessary. If the input signal is clean enough, the sound quality of the receiver will be suprisingly good. In this respect it really stands out amongst similar commercial radios.

Finally, the receiver is remarkably inexpensive. If, like countless other constructors, you have a 'junk' box full of ferrite rods, tuning capacitors and transistors, it will only cost a few pence.

noise at high frequencies

(an important factor)

Noise in UHF/VHF receivers can be determined by using extensive and expensive test equipment. However, tests with a noise generator can give usable results at a much lower cost. Such a noise generator can, of course, be constructed by the amateur.

What is noise?

Noise is caused by highly complicated physical and thermodynamic processes. Briefly, it is the random movement of electrical charge carriers. Noise increases with rise in temperature: at absolute zero (−273°C = 0 K) noise is zero, for at this temperature all movement is frozen. This is why during certain critical processes, cryogenic techniques are used to attenuate the noise factor with very low temperatures. However, it is not always practical to go to these extremes.

The signal-to-noise ratio is the best known method for determining to what extent noise (N) affects the signal (S). This can be done by expressing the signal-to-noise ratio in dB:

$$S/N = 10 \log S/N \text{ dB}$$

Taking a certain point in the receiver (after the demodulator for instance), it can be determined how many microvolts are required at the input in order to obtain a certain signal-to-noise ratio at the output.

How to determine the noise factor

The noise factor in receivers can be calculated in two different ways, either from a sensitivity or from a noise measurement. In order to test sensitivity a signal generator is required; however, good quality HF signal generators tend to be very expensive . . . Instead of measuring the sensitivity with only one frequency, we can apply many frequencies at once: use a noise signal, in other words.

This is how it works. First the basic noise N of the receiver is measured when the noise generator is switched off. Then the noise generator is switched on and the noise level is set (by means of an attenuator) in such a way that twice the input level can be measured at the output. This corresponds to a S/N ratio of 3 dB. The nice thing about using noise methods is that the S/N ratio is not dependent on temperature or bandwidth.

Circuit

A small generator can be built with inexpensive and readily available components as shown in figure 1. A high frequency transistor (T2) is connected as a zener diode. It is fed by a DC voltage source (T1). The noise voltage and therefore output level is determined by the setting of potentiometer P1 which controls the amount of current that flows through the zener diode. The output impedance of the circuit is approximately 50 Ω. The photograph in figure 2 shows part of the generator's noise spectrum.

Obviously, the circuit cannot be expected to perform miracles. The stability (temperature coefficient of the voltage source T1) achieved in the long run is not ideal, but for comparative (short term) noise tests it is quite adequate.

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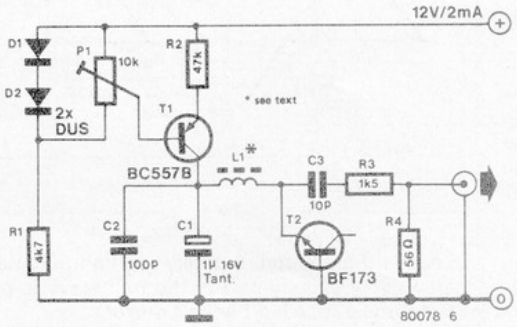


Figure 1. Circuit diagram of the simple high frequency noise generator.

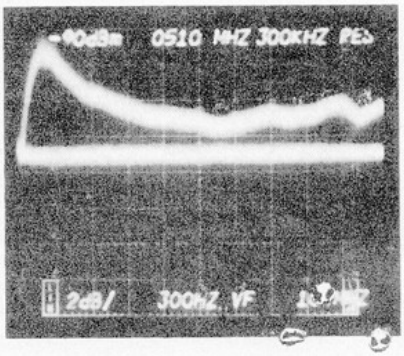


Figure 2. Part of the frequency spectrum of the circuit of figure 1. On the upper trace the frequency is shown horizontally (100 MHz per cm) and the amplitude vertically (2 dB per cm). The lower trace represents the noise produced by the spectrum analyser (−97 dBm: 0 dBm is 1 mW with an impedance of 50 Ω).

Dimmers

".... Have you noticed the advertisements for dimmers in the newspapers? There are so many of them now a days!"

"Yes, with these dimmers, we can adjust the light to be brighter or darker."

"Is there a transformer inside such dimmers to increase or decrease the voltage output?

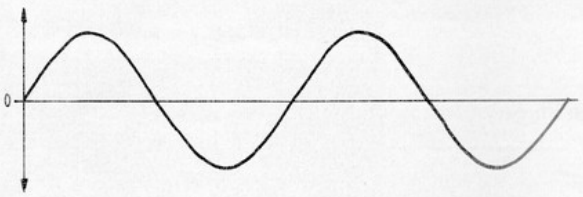
"No, there is no transformer inside the dimmer. It works electronically"

"You mean there are electronic transformers?"

"It is rather an electronic switch called TRIAC and it works on the principle of Phase Control."

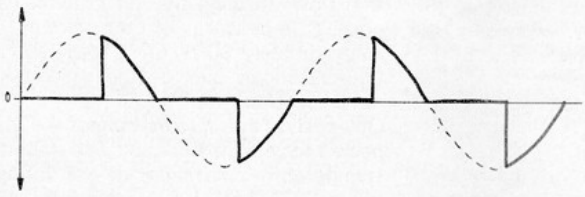
"Phase? I have heard something about the Phase. It has something to do with the alternating voltage!"

"Correct, The waveform of the alternating voltage is a sinewave. This is how it looks".



"I know this, the voltage alternately becomes positive and negative. But what does a dimmer do to this sinewave?"

"The TRIAC inside the dimmer acts as a switch and allows only a part of every half wave to pass through to the light bulb. The waveform then looks somewhat like this!"



"How does this happen?"

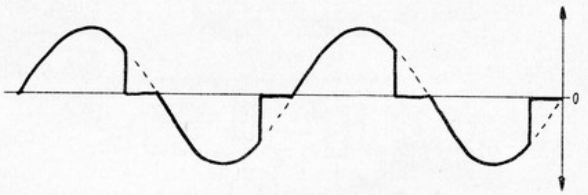
"This happens due to the properties of the TRIAC. It is a switch which can only be switched ON, and it does not switch OFF unless the current flowing through it drops to zero"

"... and the alternating voltage automatically becomes zero after every half cycle, so the current flow also must become zero every half cycle.

"Exactly, that is what happens inside the dimmer. We can switch ON the TRIAC at any place over the half wave, and it automatically switches OFF when the waveform crosses the zero level. The process repeats on every half cycle."

"But then how does the light become darker or brigher?"

"For this we just have to adjust the point at which the TRIAC is switched on or triggered. The more the half wave passes through, the brighter is the light."



Now I understand, to make the light dimmer, we must allow only a small part of the half wave to pass through. But why is it called Phase Control ?"

"Because the point of time of a wave is related to the Phase angle and the brightness of the light depends on the Phase angle of the wave at which we switch on or trigger the TRIAC."

"And that is why we call it Phase Control"

"Yes, If we trigger the TRIAC at the beginning of a half wave, we get the brightest light. If we trigger the TRIAC at the end of a half wave we get a very dim light."

"Wait a minute. It is really useless to trigger the TRIAC before tha peak valve. The maximum value will then always pass through!"

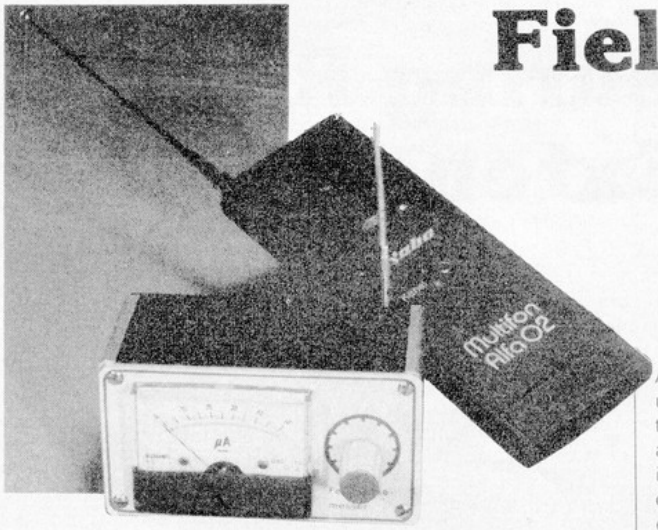
"True, the peak value will always pass through, but it is not the peak value that decides the brightness. It is the effective value that decides the brightness."

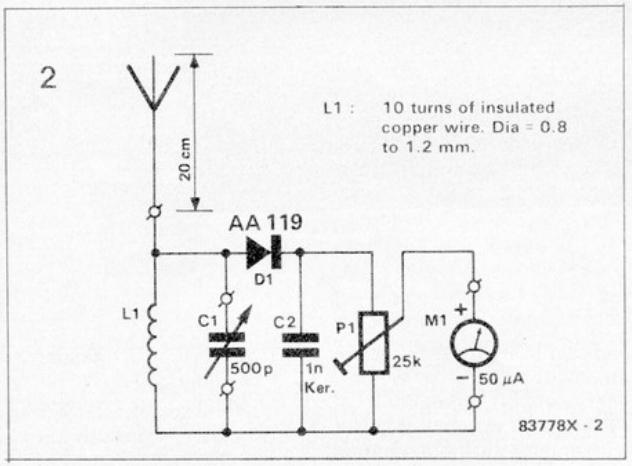
"Oh yes, I remember, the effective voltage is less than the peak value".

"The effective voltage depends upon the area of the wave that passes through"

"Then the Phase Control is nothing else but simply Effective Voltage Control. I always wonder why the electronics experts make things so complicated?"

Field Strength Meter





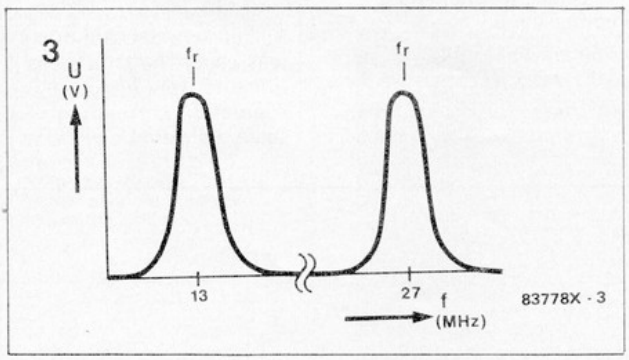


Figure 1:

The signals emitted by a transmitter scatter in free space. The farther away we go from the transmitter, the lower is their strength.

Figure 2:

The most important part of the circuit is the parallel resonant circuit formed by L1 & C1. The resonant frequency of this circuit decides the signal frequency to which the field strength meter reacts.

Figure 3:

The resonant frequency of L1/C1 combination must be as that of the signal being measured. Only then the signal amplitude is highest and canbe processed by the further part of the circuit and indicated by the meter. All other frequencies are filtered out. Both the curves show the behaviour at the tuned frequencies of 13MHz and 27 MHz.

A field strength meter is a useful instrument for all those involved in HAM activities and those involved in remote controlled model airplanes etc. For them it is interesting as well as useful to know, how much energy is being transmitted by the radiating antenna. Also, with the field strength meter, one can search for the most suitable antenna for a transmitter, and do the calibration. It can also be used for determining the transmitting characteristics of an antenna.

Everyone of us owns a field strength meter in the house: the amplifier system. Though it does not measure any transmitted field strength, it certainly detects and amplifies the 50 Hz hum. When you touch an input terminal with a finger, you can hear a loud hum coming from the amplifier. In this case, our body serves as the receiving antenna and the 50Hz disturbance is picked up and amplified. If you take the amplifier outside the house and away from it, the hum vanishes. Inside the house, there is a strong field set up by the current carrying mains wiring. Outside the house, this field dies down.

Another field strength meter we come across is the car radio. When we are passing below a high voltage line on the road, it picks up the 50Hz interference and amplifies it. Once we cross that part of the road the field set up by the high voltage overhead transmission line dies down and the hum from the radio diminishes.

The purpose of the field strength meter described here is not the measurement of any interference field. This instrument is used for determination of the field strength of some useful signals, like those of a remote control transmitter. Near the transmitter antenna, the field is very strong and then it expands in free space. The farther we go from the transmitting antenna, the broader is the scatter and weaker is the field strength. The energy available from the transmitted signal goes on reducing as we go away from the transmitter antenna.

A remote controlled model must move within the area where the field strength is enough for the model to react with. Our field strength meter can be used to decide upto what distance the field is strong enough. The field strength is normally stated in mV/m, however the device described here indicates just the ratio value, from which a conclusion about the actual field strength can be easily drawn.

The Circuit

Figure 2 shows the circuit diagram of the field strength meter. It consists of just four passive and one active component, one receiving antenna and one 0-50 µA meter movement. It does not need a supply voltage, as it draws the energy directly from the field which it is measuring. The antenna picks up the HF signal transmitted by the transmitter. The parallel

Measuring Range Extension

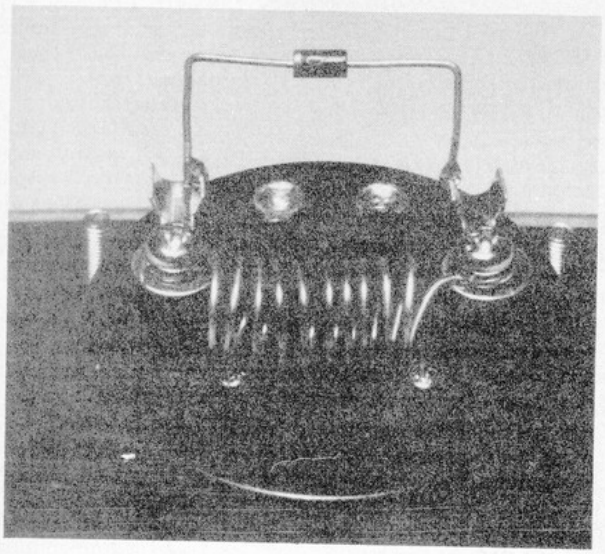
A question most frequently faced by a beginner is "I have a 100μA measuring instrument and I would like to measure 1A with it, how do I do it?" The solution is quite simple. Use a parallel resistance! In technical jargon this is called a 'SHUNT'.

Simple! isn't it?
Not quite so, there is a small problem with the shunt. The required parallel resistance can be calculated only if we know the internal resistance of the measuring instrument, and even if we are able to calculate it, it is most likely that the calculated value will not be a standard value.

A simple and practical way of making a shunt is shown in figure 2. Ordinary solder wire is used as a shunt resistance. The required length can be determined by trial and error.

The equipment required for this is one variable voltage source, one multimeter, a resistance of 10Ω/5W and some solder wire. A diode (1N 4001) can be used to protect the measuring instrument.

Now, using the multimeter, it is possible to set the current in the circuit at 1A by adjusting the supply voltage. The length of solder wire in parallel with the instrument is adjusted in such a way that the needle deflects to full scale at 1A current through the circuit.



It is also possible to set the shunt in such a way that the full scale deflection stands for 2A or even 3A. Be careful and remember to disconnect the power supply when even the solder wire is being removed for adjustments.

Results of an experiment carried out with a set up as shown in figure 2 are given below for reference which can serve as a guideline. The actual values will differ depending on the diameter of solder wire and its composition.

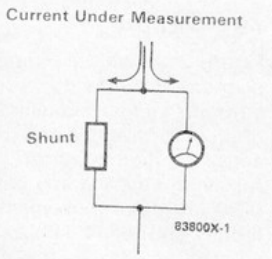
The typical results are as follows :

Instrument 100μA F.S.D. movement.

Modified Range	Solder Wire length
1A	34 cm
2A	16 cm
3A	10 cm

Solder wire of 0.8 mm dia was used. The shunt was found to work effectively even at 3A current with full scale deflection.

1



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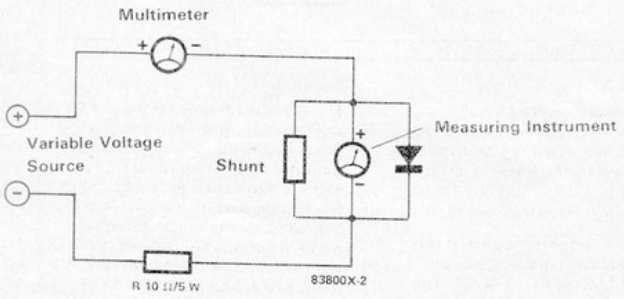

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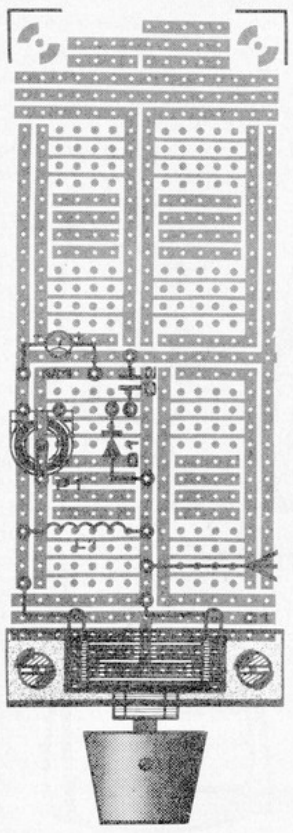


Component List

C1 = 500 pF, Variable
C2 = 1 ηF Ceramic
P1 = 25 KΩ, Potentiometer
L1 = 10 turns of insulated copper wire (dia 0-8 to 1.2 mm)
D1 = AA 119 (Germanium)
M1 = Moving Coil meter 0-50 μA.

Other parts

1 SELEX PCB
1 Antenna
1 Suitable enclosure
3 Banana plugs and sockets if necessary.

Figure 4:

The component layout on a standard SELEX PCB.

Figure 5:

The photograph shows internal construction of the field strength meter with the back cover removed.

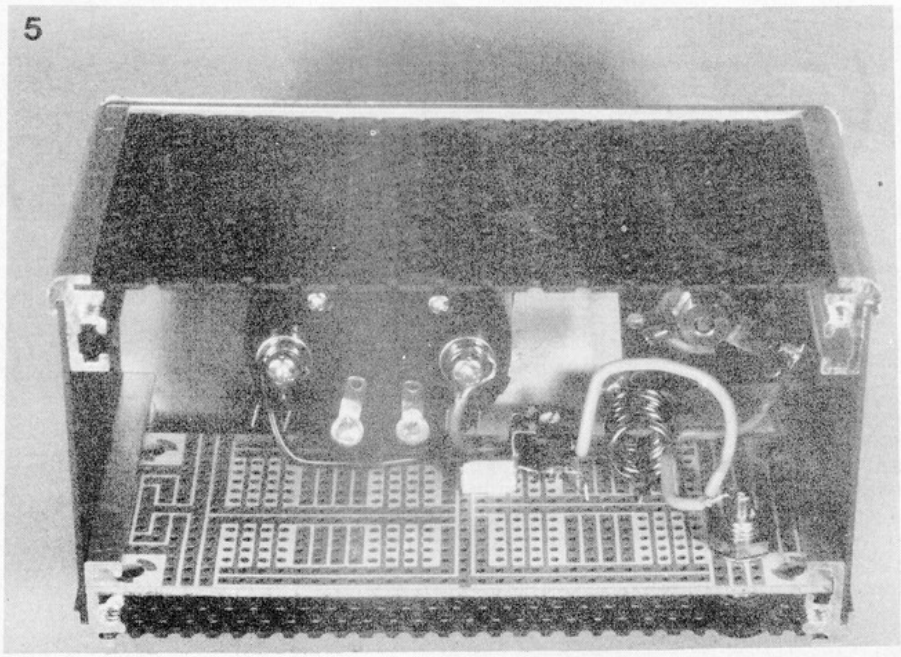
resonant circuit, made up of coil L1 and the capacitor C1 is used for tuning to the frequency of the signal. Capacitor C1 is a variable type, with a continuous adjustment possible between zero and maximum value. With the given values of L1 and C1 it is possible to tune to any signal from 13 MHz to 40 MHz. This range covers both the radio amatures as well as the remote control enthusiasts.

Along with the antenna, the parallel resonant circuit picks up the signal to which it is tuned and passes it on to the indicating part of the circuit consisting of the diode D1, capacitor C2, potentiometer P1, and the meter M1. The amplitude of the signal being picked up is maximum at the tuned resonant frequency fr. Figure 3 shows this graphically for 13 MHz and 27 MHz. If C1 is moved away from the tuned position, the amplitude suddenly drops down. The tuning of the parallel LC circuit depends on the resistive component of the impedance of L and C. With increasing frequency the resistive component of Coil impedance increases, whereas that of the capacitor impedance decreases. The circuit resonates at the frequency at which the resistive components of both the coil and capacitor impedance become equal to each other As in our circuit the coil L1 is fixed and C1 is variable, the change in resonant frequency is entirely dependant on the value of C1. The signal is rectified by the diode D1 and a DC voltage is generated across C2. Potentiometer P1 converts it into suitable current for the meter M1 to indicate the value. Setting of P1 decides the sensitivity of the field strength meter. The sensitivity should be adjusted depending on the field strength being measured. The indicator deflection depends on four factors: the transmitter output capacity, the antenna, the distance between transmitter and the instrument and the sensitivity setting of P1.

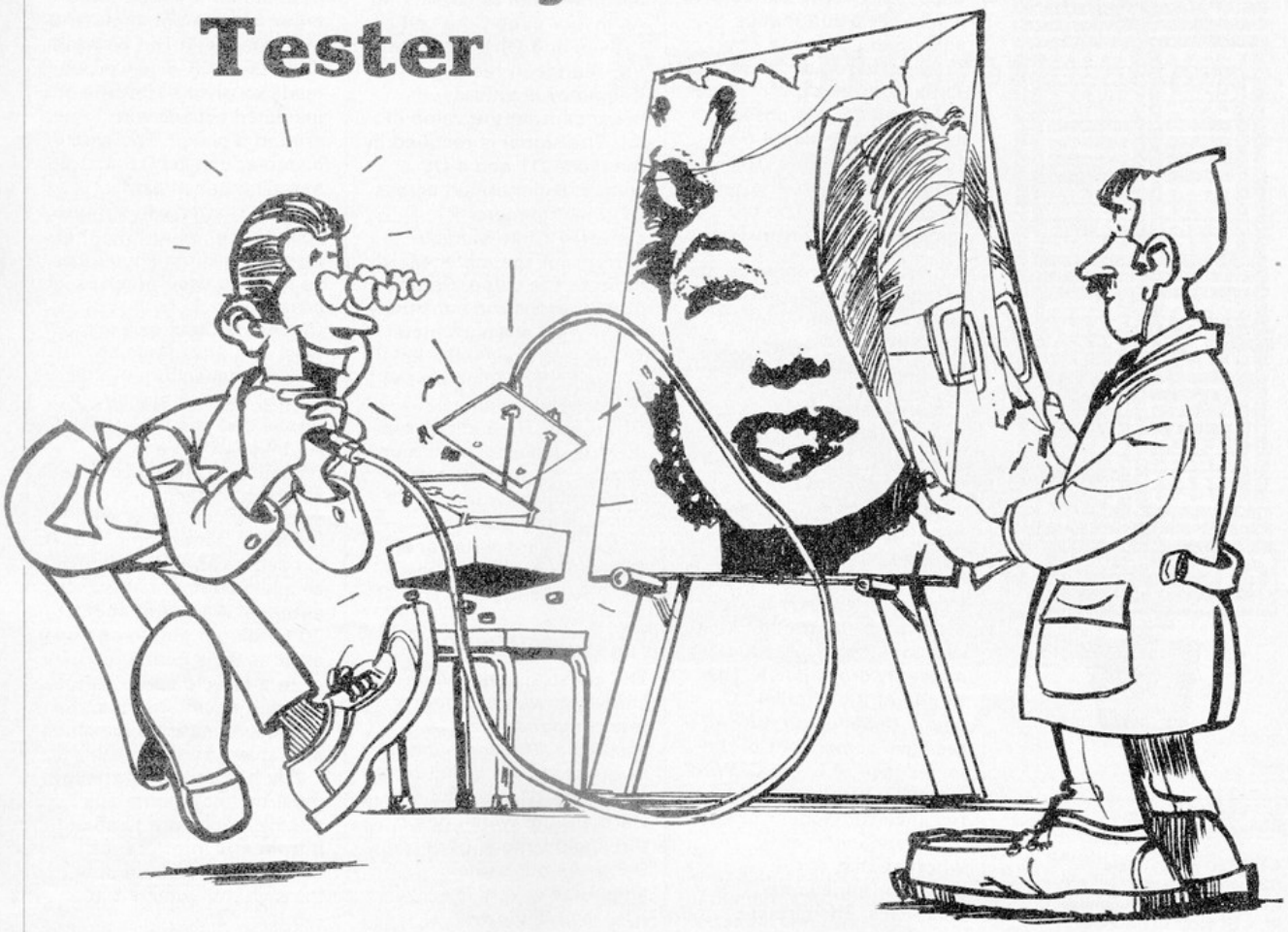
The Construction

The construction is very simple as we have only a few components to assemble. The component layout is shown in figure 4. The diode D1 must be of Germanium, which has a threshold voltage of only 0.3 V. As our signal amplitude is very low, a Silicon diode with a threshold of 0.7 V will not be suitable The variable capacitor can either be mounted directly on PCB if possible, or it canbe fitted on an aluminium angle and then the angle can be fixed on to the PCB. The Coil L1 is made by giving 10 turns of insulated copper wire around a pencil. The wire diameter can be 0.8 to 1.2 mm. The pencil can be taken out after winding the coil without disturbing the turns. The turns should be as close to each other as possible.
C2 should be of ceramic type. The advantage of ceramic capacitors in that they have high stability, low losses and good tolerance. All these affect the measuring range of the field strength meter.
A moving coil meter with a full scale deflection of 50μA is used as an indicator. The simplest component is the antenna. A wire of about 20 to 30 cm length can be used as an antenna, or even a bicycle spoke can be used as an antenna, if a telescopic antenna is not available.
Lastly, the entire instrument must be enclosed in a metalic enclosure to shield it from any interference fields and to pick up only the signal to which it is tuned.

5



Continuity Tester



With the last pin soldered, the soldering iron kept aside, we are finally ready with our PCB with all components assembled! We connect the power supply and switch it on... nothing happens !!

Unfortunately, after so much of effort we dont get the result. At this point what we badly need is a continuity teser. Because, if the circuit does not function after we have asembled it correctly, most probably these is something wrong with the connections. Either a soldered connection is not good enough or the solder has unintentionally spread to unwanted areas - creating a short circuit where no connection was desired. Or, may be we have missed a connection totally and left it unsoldered.

All these possibilities can be checked with a simple continuity tester described in this article. Connections between components can be quickly checked with this continuity tester. An audible beep indicates that the connection is proper. Absence of beep shows an open connection, or a high contact resistance. If the tester beeps when connected to two points which are not intended to be connected to each other, it means there is an unwanted short circuit.

The tester circuit is so designed, that a contact resistance of more them 1 Ω shows as an open circuit. There is no risk of damaging the sensitive components by the tester because the test signal is very small.

The Circuit

The main component of the tester circuit is the Operational Amplifier (Op Amp) IC1. The Op Amp is connected as a differential amplifier, which means that the amplifier reacts only to the difference between the voltages on the inverting (Pin 2) and the non inverting (Pin 3) inputs. The difference signal is strongly amplified and carried to the output (Pin 6). Normally the points A and B are at equal voltage level, namely at approximately half the operating voltage. This is so because R1 and R3 have equal values. The introduction of R2 in the circuit makes the voltage at A more positive by about 2 mV compared to point B. Thus the input pin 2 lies at somewhat higher voltage than pin 3. As the Op Amp has a very high gain,it switches the output to zero volts.

Now if the test tips are connected to each other, R2 gets short circuited. The potentiometer P1 is so adjusted that when R2 is short circuited, the voltage at pin 3 is slightly more positve than that at pin 2. This gives a positive voltage at the output of the Op Amp.

The Op Amp behaves in the following manner:

In case of short circuited measuring tips, the output goes high, approximately to the supply voltage. If the test tips are connected across a resistance of more than 1Ω, then the Op Amp output becomes zero.

To convert the output voltage into an audible output signal, an oscillator is constructed with N1, R7, P2 and C1. The oscillator output drives the piezo-buzzer, Bz. The oscillator is an astable multivibrator and produces a 4.6 KHz signal. The current taken by the oscillator is about 3 mA, to drive the buzzer.

The frequency can be adjusted with P2, and the frequency adjustment also affects the volume because the buzzer works most efficiently at its resonant frequency. The oscillator is turned on or off by the Op Amp output signal. When the output is "0", there is no audio output. When it becomes "1" the oscillator starts driving the buzzer and produces sound output.

The oscillations take place as the capacitor C1 is alternately charged and discharged. This produces a square wave, which drives the Piezo-Buzzer.

Construction & Adjustment

Because of the few components used, the construction is quite simple. A component layout on SELEX PCB is shown in figure 2. Correct polarity must be observed for C2, and Pin numbers of ICs must be properly connected.

A 9V miniature battery is used as the power supply and connected through the switch S1.

The assembled circuit must be tested before fitting into the enclosure. For this purpose, P1 and P2 both should be kept in their center position. On shorting points A and B, the buzzer should give an output. If buzzer does not make any sound, adjust P1 till it just starts making a sound. Then adjust P2 to set the volume. For correct adjustment, a 1Ω resistance is connected across A and B and P1 is set at a point where the buzzer just starts giving an audible output. In this way the tester will not indicate a short circuit even for small values of resistors.

If nothing works as outlined above, check the PCB assembly carefully. Are all the components correctly placed? Are they all of correct values? Also check the voltage at A and B, it should be approximately about 4.5V on both points. The output of IC1 must be low when terminals A and B are not shorted. It should be almost equal to the supply voltage when A and B are shorted.

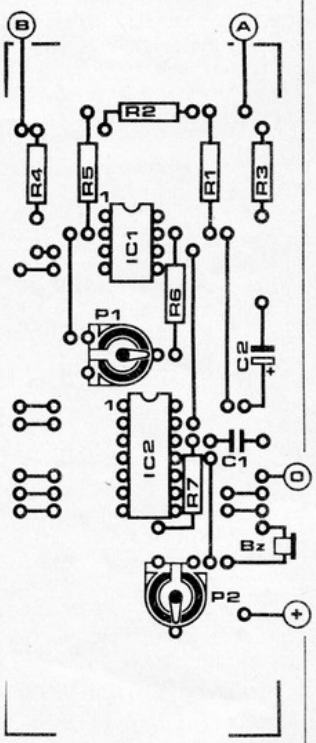
If all this is in order and still the tester does not work, check the wiring.

Once the circuit starts functioning, it can be fitted into a small handy plastic cabinet. Such an assembly is shown in figure 3.

R1 R3 = 22KΩ
R2 = 10Ω
R4, R5, R7 = 1KΩ
R6 = 470 KΩ
P1 = 10KΩ Trim Pot.
P2 = 2.5KΩ Trim Pot.
C1 = 100 nF
C2 = 10 µF/10V
IC1 = 741 OpAmp
IC2 = 4093

Other Parts :

S1 = ON/OFF Switch
1 Battery clip
1 9V miniature Battery
1 Piezo Buzzer
4 PCB terminals
1 SELEX PCB
1 Suitable enclosure
Flexible wire and Test Tips.



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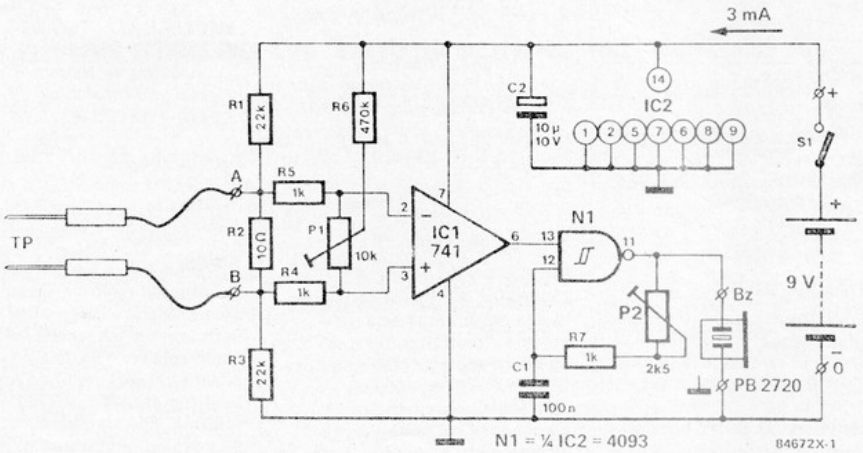


Figure 1:
The continuity tester indicates short and open circuits in an assembled electronic circuit board. For resistances less than 1Ω it indicates a short circuit by giving an audible beep sound.

Figure 2:
The entire circuit of the continuity tester can be mounted on a SELEX PCB. Only the sockets, switch and battery are fitted externally.

The installation in a plastic enclosure is quite easy. All one has to do is just drill a few holes and fit everything inside.

NEW PRODUCTS

ELECTRONIC WEIGHING SCALE

AFCO have developed an Electronic Precision Weighing Scale, which works on the strain guage Loadcell.

Having a capacity of 1999 gms it gives a reading on a LCD screen with ± 1 gm resolution. The scale features a detachable deep square pan and a rotary switch to correct the reading when the deep pan is not in use. Compact in size, it is suitable for use in office, Kitchen, laboratory, postal department etc.

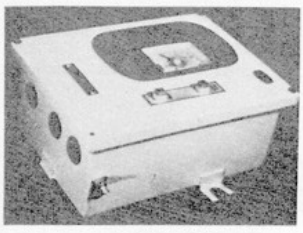
For further details, please contact.

M/S AFCO MARKETING PVT. LTD.
104 Creative Ind. Estate
72, N.M. Joshi Marg
Sitaram Mills Compound
Bombay-400 011.

Phone: 39 89 34/89 64 72

WINDING TEMP. INDICATOR.

PECON has developed a solid state winding temperature indicator for power transformers. Available from 2 to 4 contacts with repeater and 4-20 mA output for DAS operations, it can withstand all types of voltage and current surges prevalent in power plants. This indicator comes for various voltages upto 250V in both AC & DC.



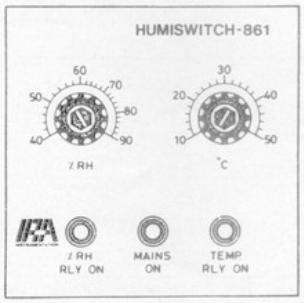
For further information, contact.

POWER ELECTRONICS & CONTROL,
612, Yashkamal,
Tilak Road, Vadodara-390 005

HUMIDITY SWITCH:

IRA has developed a humidity switch, the Humiswitch - 861 for control of humidity in instrumentation, LT & HT control panels etc. This switch, based on LSI technology is reliable, accurate, compact and useful for controlling humidity to tolerable limits to avoid the risk of flashover and mal operations.

Humiswitch 861 has an electronic transmitter and an electronic control switch connected by a 5 core cable. The electronic transmitter is equipped with sensors & signal conditioners for both RH and dry bulb temperature and can be used for controlling only humidity or both humidity and dry bulbs temperatures. Where temperature control is required, a circuit for giving alarm/switching on blowers or coolers can be provided for keeping the temperature within reasonable limits.



For further details, contact.

M/S INSTRUMENT RESEARCH ASSOCIATES PVT. LTD.
P.B. No. 2304/1-2 Magadi Road
Bangalore-560 023.

DIGITAL MEGGER.

SCR Electroniks have developed a digital megger, Digi Meg. Available in two basic versions viz. LED model (suitable for 230V AC or 12V DC) and LCD model (working on 9V battery) the models feature - range of upto 2000 meg ohms, resolution of 0.01 meg ohms and 0.1 meg ohms respectively, accuracy of ± 1%, test voltage of 500V DC and 3½ digit 0.8" display (LED or LCD)

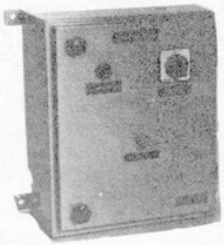
For further details, please contact.

SCR ELECTRONIKS
Opp: Fatima High School
Kirol Road, Vidyavihar
BOMBAY-400 086.

AMPLITUDE CONTROL

Static Power Systems have developed solid state amplitude control panel for electromagnetic vibrating feeders, screens and other equipment. This control features a smooth start from zero to the full capacity, plug-in-type modular cards, voltage variation, isolation transformer for control & trigger circuits. etc.

These controllers are available upto 500 V DC and 70 amps.



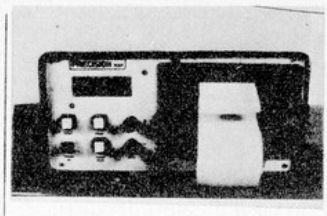
For further information, please write to:

M/S STATIC POWER SYSTEMS PVT. LTD.
D-148 Bonaza Ind Estat
Ashok Chakravarty Road
Kandivli (W) Bombay 400 101

PHONE: 691173

D C CALIBRATOR

PRECISION'S Universal D.C. Calibrator is a microprocessor based instrument which can measure the output of any standard thermocouple type B,E,J,K,R,S,T. It features automatic cold junction compensation, temperature display with a resolution of 0.1°C, measurement as well as simulation of millivolt and milliampere signals etc. The unit comes in a portable cabinet with simple panel controls.



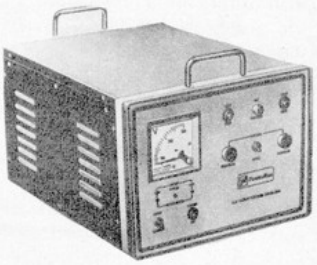
For further information contact:

M/S PRECISION ELECTRONICS
38-D GIDC Bhaktinagar
RAJKOT-360 002 (Gujarat)

VOLTAGE STABILIZERS

Digitech have developed the 'Powerline' servo controled a.c. voltage stabilizers which stabilize the voltage to 230V ±0.5% in 1-0 and 400V ± 0.5% in the 3-0 version. A number of models rated from 0.5 KVA to 100 KVA are available. The control circuit is I.C. based and 3 LEDs on the panel indicate the input line status and the correction being carried out. The correction rate is 30V/Second

Other features iclude: No waveform distortion, no rf interference, output voltage adjustability, auto/manual mode of operation and voltmeter for voltage monitoring.



For further details, please contact.

DIGITECH ELECTRONICS
D-1/GF-9 Virat Bhavan
Dr. Mukherjee Nagar
Commercial Complex
DELHI-110 009

SMPS

Omnitronix have introduced switch Mode Power supply with built in automatic voltage stabilizers for TVs. Mains isolated, it features line and load regulation, EMI/RFI filtering, line transient supression circuit and a modular construction.

For further details, contact:

M/S. OMNITRONIX
C-5, Amijyot, Ambawadi,
Ahmedabad-380 006.

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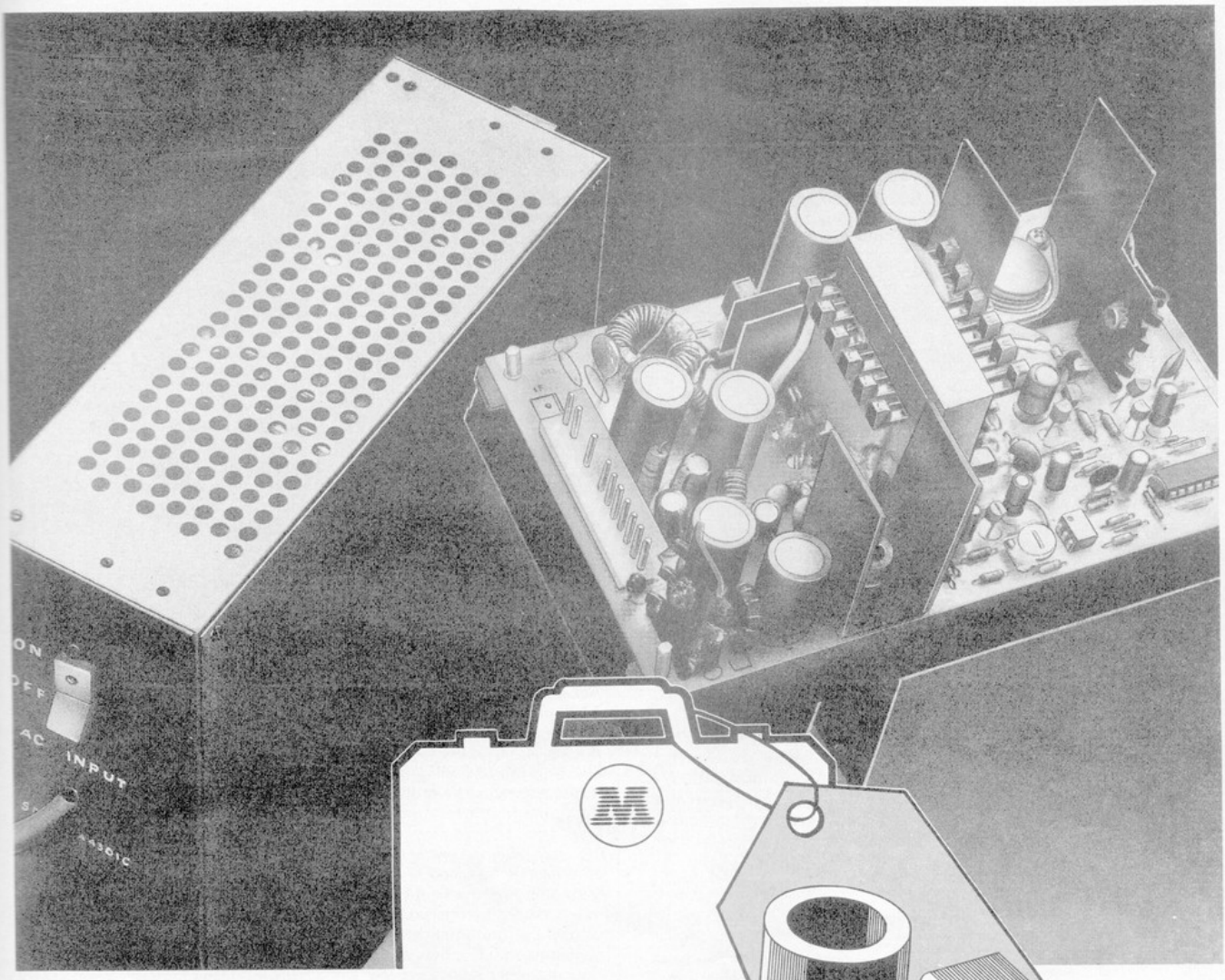

Murugappa Electronics Ltd offers you high quality soft ferrite cores, manufactured by Hilversum Electronics – a name to reckon with

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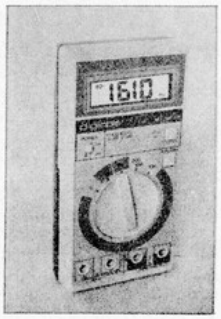
LINTAS M MEL PPG 72719

NEW PRODUCTS

Digital Multimeter

Ledtron have introduced Digicone 1210 autoranging Digital multimeter. Measuring upto 1000V AC/DC; 20 Mohms; 10 A AC/DC the multimeter can test diodes and has a buzzer for audible continuity testing. It has a rugged design, is portable and compact.



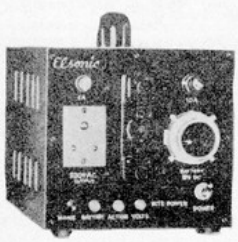
For details contact:

M/s. LEDTRON ELECTRONICS,
170 Lohar Chawl,
Bombay-400 002.

UPS

Elsonic have developed Nitepower 200 SN solid state uninterruptible power supply system. Sometimes referred to as a storage type mini generator, the system is compact and generates quare waves at 50Hz which are fed to the output triacs which turn on only when the mains fail.

The unit gives 160W of power during a power failure for 3 hours when connected to a 12V, 88Ah rechargeable lead acid battery. The Nitepower has a maximum switching time of 5 seconds and is suitable for operating lights, fans, telex machines, telephone exchanges, intercoms, TVs and computers.



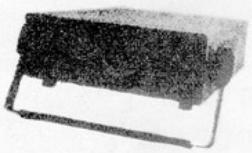
For further details contact:

ELSONIC SANTO CORPN.,
8/1, Palamgrove Road,
Bangalore-560 047.

LCR meters

Murugappa Electronics Ltd. are marketing AG-4301B and AG-4303 digital LCR meters made by Ando Electric of Japan Designed for measuring components quickly and accurately, the meters use a new measurement technique and displays L, C, R, D and Q at 2 different frequencies of 1KHz and 100 Hz. In addition it also has a residual charge protection upto 50V, facility for superimposing upto 30V DC.

If used in conjunction with Antlo's comparator AG4902, fast GO/NO-GO determination of components can be made against preset upper and lower limits of component parameters.



For details contact:

MURUGAPPA ELECTRONICS LTD.
Agency Division, TIAM house
28 Rajaji Road,
Madras-600 001.

Digital Timers

Alfa have developed a range of Timers with quartz crystals. Having a high timing accuracy of 0.01% the timers have a wide range of operation.

The time setting is done through thumb wheel switches and the set is displayed on the switches. When the timer is switched ON, the LCDs display the count down and as the time lapses, the display indicates a zero and an internal relay energises to changeover the contacts. A push button switch is provided for resetting the time.

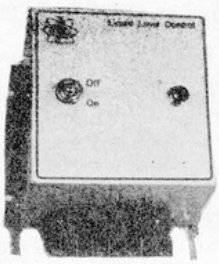
The contact is rated at 230V/6A and the timers are available in Delayed ON or Delayed OFF models. The models come in 2to 4 digit display and blind versions.

For further information contact:

M/s.ALFA PRODUCTS COMPANY,
FF-11 Bajan House, 97 Nehru Place,
Post Box 4324
New Delhi-110 019.

Liquid Level Controller

Accents Liquid level controller (LLC) monitors the level of a liquid in a tank. It ensures that this level either does not rise above or fall below some preset thresholds, thereby protecting the pump from over/under load. The LLC has a compact 10cms x 10cms control unit with probes attached. Depending on the liquid level, as measured by this probe, the control unit automatically switches on/off either a pump or some other device whose operation is to be controlled

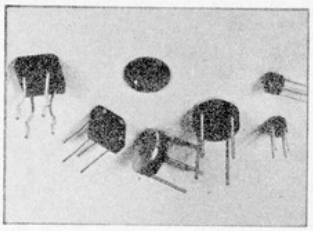


For further details, contact

M/S. ACCENT CONTROL (P) LTD.,
Kataria Mansion, Dr. Annie Besant Road,
Worli, Bombay-400 018.

Bridge Rectifiers

Solid State Electronics have developed a series of light weight efficient, full wave single phase rectifier bridge. Available in two types, PCB mounted or chassis/heatsink mounting types, they are rated for currents of 1 to 8 Amps for a voltage range of 6Vrms to 700Vrms.



The internal construction has a low thermal resistance and low forward voltage drop. The rectifiers can be used in electronic instrumentation, controllers, consumer electronics, business machines, power supplies etc.

For further details, write to

SOLID STATE ELECTRONICS CO. PVT. LTD.
Plot No. 9/123, Marol Co-op. Ind. Estate,
P.O. Box. 7432, J.B. Nagar Post,
Bombay-400 059.

Electronic Pain Killer

Johari have introduced an electronic pain killer based on Transcutaneous Nerve Stimulation therapy. It removes the pain from the nervous system and muscles by sending external electrical energy to the affected parts.

The instrument work on a 9V battery, weighs only 75 gms and comes in a range of pleasing colours.

The device is useful in backache, migraine, arthritis, muscular pain, sprains and fatigues.

For further information contact:

M/s. JOHARI ELECTRO-TECH CO.
Vandana, 28, neharu park,
Jodhpur-342 003.

MOTOROLA

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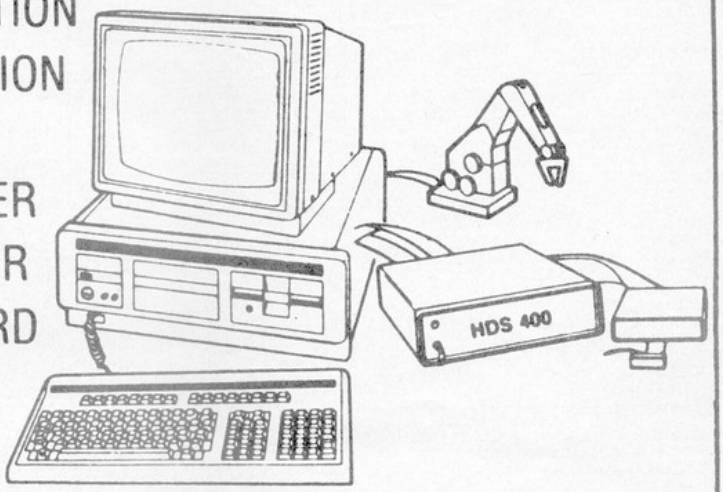
* OEM USER STATION
* SINGLE BOARD COMPUTER
* PERIPHERAL CONTROLLER
* A/D D/A INTERFACE BOARD
* CARD CAGE, BACKPLANE
* CHASSIS AND POWER SUPPLY



Semiconductor Products

INTEGRATED CIRCUIT, RECTIFIER, ZENER, SCR, PRESSURE SENSOR ELEMENT, UJT, FET, PHOTOELECTRONIC PRODUCT, DISCRETE MICROMINIATURE PRODUCT.

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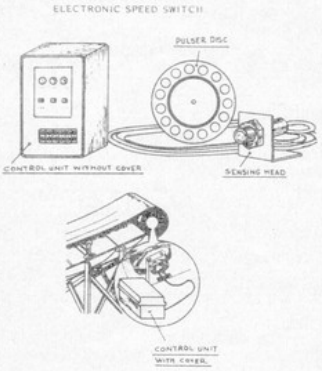

For further information contact:

M/S. INDIAN ENGINEERING COMPANY
Post Box 16551,
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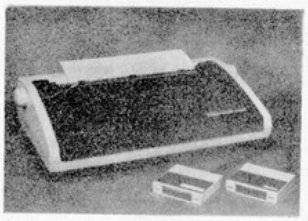

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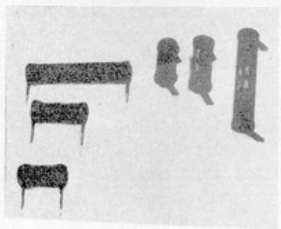

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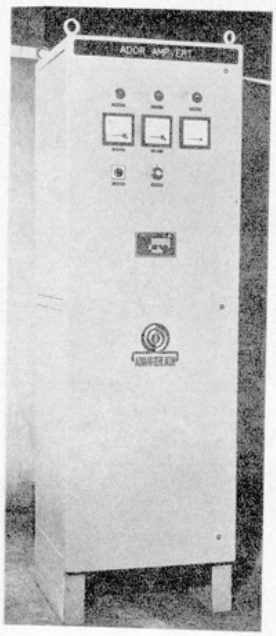


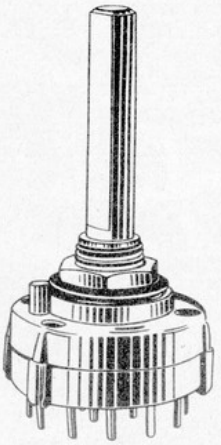
For more details, quote ref. No. P3/2/84 and write to :

Advani-Oerlikon Limited
Post Box No. 1546
Bombay 400 001.

ROTARY SWITCH WITH ADJUSTABLE STOP TRS-12

'Comtech' TRS-12 is a Rotary switch which is offered in 3 models, FirstL 1 pole 12 positions, Second: 2 pole 6 positions and Third: 4 pole 3 positions. A stop ring is provided with each switch to enable the user to adjust the 'switch-stop' position to meet his requirements. Each model is available with printed circuit and solder terminals having tin-glow plating to impart quick solderability.

The body is made up of glass-filled Nylon and the contacts are made from brass and phosphor bronze. The switch at present is offered with nonshorting contacts (Break before make), having a rating of 350 mA at 110 V Ac/Dc, and with contact resistance of 20 milliohms max.



The switch is having a minimum mechanical life of 20,000 detent operations. The shaft accepts any knob suitable for 1/4" diameter.

For futher details contact:

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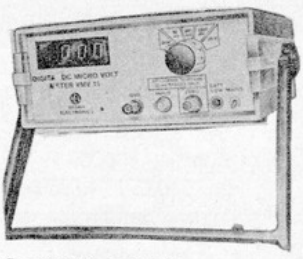

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'S' correction aluminium electrolytic Capacitors.
– an import substitute from ELCOT.

RANGE

MFD	2.2	4.7	6.8	10	12
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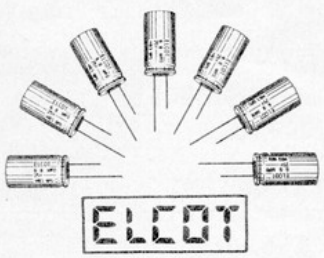

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CORRECTIONS

Low noise aerial booster
(Aug Sept 1986-p. 21)

On the component overlay, the collector of T_1 is erroneously shown to be connected to the PCB ground plane, while it should go to junction L_2-C_4-R_4-L_3 (turn the transistor 90° cw).

Top of the range Preamplifier (Part 1)

(Dec 1986 p. 12-24)

The infocard mentioned in the note on page 12-24 (see note printed in Italic) will be published in the February 1987 issue.

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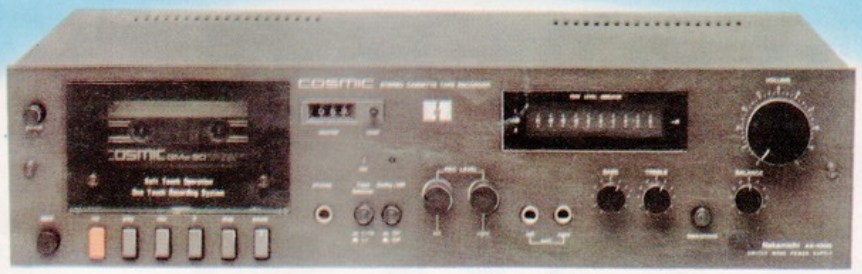

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